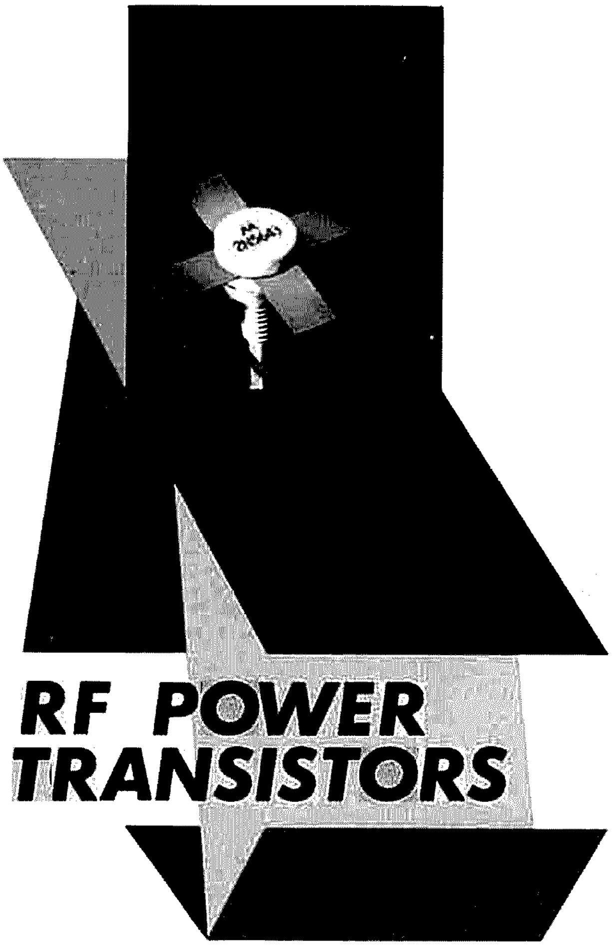
75 cents

hr

focus on communications technology . . .

# ham radio magazine

JANUARY 1970



## RF POWER TRANSISTORS in amateur transmitters

### this month

january 1970
volume 3, number 1

## staff

James R. Fisk, W1DTY
editor
Nicholas D. Skeer, K1PSR
vhf editor
J. Jay O'Brien, W6GDO
fm editor

A. Norman Into, Jr., W1CCZ
Alfred Wilson, W6NIF
James A. Harvey, WA6IAK
associate editors

Jean Frey
art director
Wayne T. Pierce, K3SUK
cover

T. H. Tenney, Jr., W1NLB
*publisher*

## offices

Greenville, New Hampshire 03048
Telephone: 603-878-1441

**ham radio** magazine is
published monthly by
Communications Technology, Inc.,
Greenville, New Hampshire 03048

*Subscription rates, world wide:*
one year, $6.00, three years, $12.00.
Second class postage
paid at Greenville, N.H. 03048
and at additional mailing offices.

Foreign subscription agents:
United Kingdom:
Radio Society of Great Britain,
35 Doughty Street, London WC1, England.

All European countries:
Eskil Persson, SM5CJD, Frotunagrand 1,
*194 00 Upplands Vasby, Sweden.*

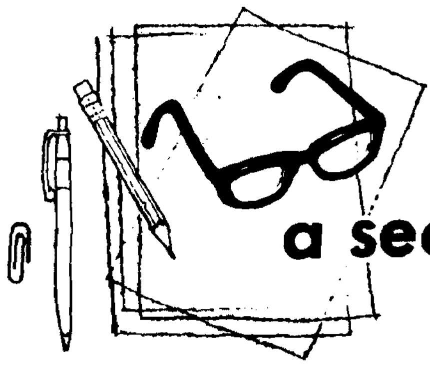

Printed by Capital City Press, Inc. in
Montpelier, Vermont 05602, U.S.A.

**ham radio** is available to the blind
and physically handicapped on magnetic tape
from Science for the Blind,
221 Rock Hill Road, Bala Cynwyd,
Pennsylvania 19004.
*Microfilm copies of current*
and back issues are available
from University Microfilms,
313 N. First Street, Ann Arbor,
Michigan 48103.

Postmaster: Please send form 3579 to
**ham radio** magazine, Greenville,
New Hampshire
03048.

**ham radio** *magazine*

# contents



# a second look

by jim fisk

**Several months ago** I mentioned briefly one of the time-proven tactics for success in working DX—listen, listen, and listen some more. It's still good advice. But after you listen, then what?

*With the DX-contest season approaching,* I thought it would be worthwhile to pass along some words of wisdom gleaned from successful contest operators with average equipment but above-average operating knowhow. It's difficult to compete against high power and elaborate antennas, but it **can** be done. The formula is simple: sharp operating and a little luck. The idea is to refine the first so the second will be a free bonus.

Consider the inevitable pile-up of stateside stations trying to work a choice DX station. The DX station will be making contacts at a tremendous clip. Often he won't send his call for 15 or 20 minutes. To compound the problem, many U. S. hams will work him and **they** don't send his call either—takes too much time. Not only is this senseless, it's illegal for U. S. hams.

If you're with it, you'll note the DX station's frequency, move on and work others, then check his spot at frequent intervals. Sooner or later his identity will be made public. Meanwhile, the competition will have wasted precious time and contest points frantically trying to identify the mystery station.

The next step is to plan your strategy so you can leap in, latch on, and leap out. This is by no means as easy as it sounds. Here are some ideas that will help.

*Note how the DX station answers calls.* Perhaps he replies to stations clustered above and below his frequency. Or maybe he works stations who transmit a few kHz higher each time. After a few contacts, a pattern will emerge. For example, how often does he answer stations on the low side of his frequency before he changes to those on the high side? By playing the law of averages, eventually you'll score.

If all seems useless despite your efforts, the wise thing to do is to move on and work other stations. Later you can return to the original pile-up, which will probably have diminished, and you can try again. Let the high-power fellows knock themselves out trying to make one difficult contact while you're busy racking up points.

## obtaining confirmations

Trying to smoke out QSL cards from foreign hams has always been a problem plaguing the DX operator. Here's an idea used by one enterprising DX enthusiast to increase QSL returns. Raised aluminum letters, such as those used for signs and house numbers, are arranged on a colored card. The card is then sprayed with a contrasting color of lacquer. Before the lacquer sets up, the card is decorated with multicolored glitter (the stuff used by stores for display ads). After the lacquer dries, the letters are flipped off. Result: a unique two-color QSL with call letters in silhouette.

Anything homemade seems to appeal to foreign amateurs. If you can boast a personalized QSL, as in this example, your returns on confirmations should increase.

***Jim Fisk, W1DTY***
**editor**

# a word from the publisher

**As we move into** volume III of *ham radio* it is a good time to stop for a minute and look back at the 22 months behind us and the future that lies ahead.

In many ways a magazine is much like a person. In its early years it is affected by the personalities and ideas of its parents. It is apt to change during this period as these parents try out new ideas in the up-bringing of their new offspring. Some work out—some don't. We can cite several of each.

As the youngster begins to move out into the world new influences begin to have their shaping effect. The subject tries to accommodate some; others seem less desirable and are avoided. You, our readers, have provided many of these influences and *ham radio* has tried to react to the many letters and suggestions which we have received. Some we have followed, while others have had to be discarded as they did not fit into the overall plans we had in mind.

One change which we're introducing this month is our new binding. This has been done mainly to overcome complaints which we have received about the magazine coming apart after going through the mail. The newer binding is a stronger one which should eliminate this problem once and for all.

None of your ideas have gone unheeded, however. We always appreciate hearing both pros and cons of *ham radio*. In fact, we must hear them if we are to do our job properly. If this magazine is not answering your needs, then it is failing in its basic mission. We don't feel this is the case, but it's up to you to let us know how we're doing.

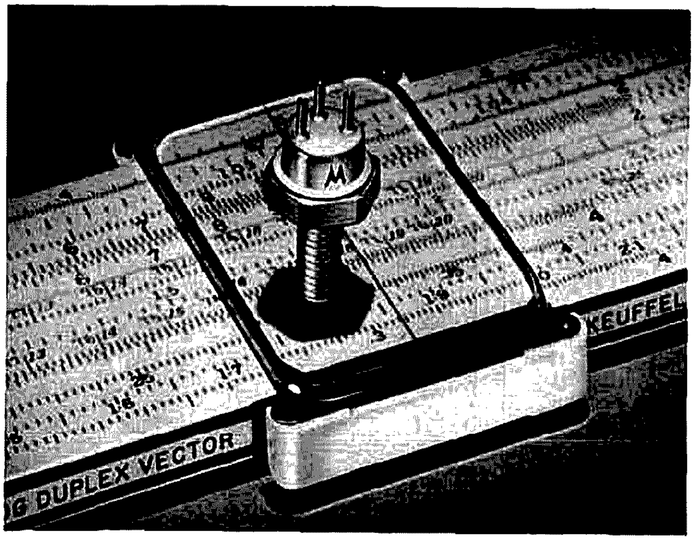
Finally, our hero begins to move out in the world and find himself. Much of this time may be spent looking for challenges. One such opportunity recently presented itself, and *ham radio* was quick to accept. We have taken over the subscription obligations of *FM Magazine* which until recently was very capably published by Mike Van Den Branden, WA8UTB. All *FM* subscribers will receive a like number of issues of *ham radio* for the months they had remaining of *FM*. Those who were subscribers to both magazines will receive an appropriate extension of their *ham radio* subscription.

This means a boost of several thousand subscribers for which we are justly thankful; but it also means much more. It means that we now reach more of the serious vhf fm operators than any other publication. *Ham radio* wants to treat this responsibility properly. With the help of Jay O'Brien, W6GDO, our fm editor, and Mike Van Den Branden, we will try to offer some really good fm material to keep our new readers up to date and introduce our older readers to this rapidly growing segment of our hobby. You can help also by lining up some good articles of this type for us and by letting us know what you would like us to print. Remember that *ham radio* pays on acceptance for any material we use.

Please bear with us for a month or so as we have not really had a chance yet to allow for this new responsibility. Lead times being what they are, it will be another month or so before we can line up some good articles and get them into print.

*Another good opportunity presented itself* a while back when we were offered the opportunity to sponsor one of those fabulous free cocktail parties at the SAROC Convention to be held in Las Vegas, February 4 through 8. Our night will be Thursday, February 5 and we hope you'll be there. Remember, this is an amateur radio show unlike any other. If you've never been to SAROC you owe it to yourself to go this year. If you've been before I know you'll be back. It's at a new hotel this year and SAROC Chairman, Len Norman, W7PBV assures us that it will be bigger and more exciting than ever before: full details are in the SAROC ad on page 78. See you in Las Vegas.

**Skip Tenney, W1NLB**
**publisher**



# how to use rf power transistors

A guide to the practical use of rf power transistors in amateur radio equipment, including circuit design, matching networks and construction

Paul Franson, WA7KRE, 7430 East Minnezona Avenue, Scottsdale, Arizona 85251

Ever since transistors were announced many years ago, hams have been interested in using them in all types of equipment. Though the advantages of transistors have made them popular for all electronic applications, transistors are especially suitable for portable equipment. Transistors, with their high efficiencies, small size, low heat dissipation, low voltage operation and high reliability, are ideal for portable gear. Many low-power transistor transmitters have appeared in ham magazines, and every circuit of this type that has appeared has attracted considerable attention. Unfortunately though, many of the transistors used in these transmitters are really not ideal for this use, since they are switching transistors or low-power amplifiers that don't perform very well in rf power service. Higher power transmitters using rf power transistors have rarely been described, and most of the circuits that have appeared really couldn't be considered very practical for most ham use.

Old rf power transistors suffer from four major faults that have limited their usefulness: low gain, limited power output, high cost, and perhaps most discouraging, susceptibility to destruction due to mismatch or detuning. This last was especially bad in mobile applications where parking too close to a vertical pipe or having your antenna touch a tree could blow out an expensive power transistor if you happened to be transmitting at the time. Various complex schemes were developed to prevent this from happening, but most were not completely satisfactory. Fortunately, new transistors overcome most of these faults.

As you probably realize, the market for transistors in amateur equipment is miniscule compared to the market in the mobile communications equipment used in police cars, ambulances, taxicabs, and so forth, not to mention the transmitters used in aircraft and military equipment. However, the amateur benefits from the improvements that result from developing new transistors for these applications. Because these markets are large and growing, transistor manufacturers have been developing highly improved transistors for these uses.

These new power transistors have higher

gain and higher power output than earlier devices (up to 100 watts in one transistor at 175 MHz). They are also rugged and can withstand detuning and mismatching that would destroy earlier devices. Their cost is reasonable for the applications they are intended for. While prices are still high compared to vacuum tubes which can supply the same power, the advantages of transistors have made them the overwhelming choice in new applications. Very little new communications equipment for mobile use is presently being designed with vacuum tubes.

For applications that require high efficiency, small size, and high reliability transistors are used even when they are quite a bit more expensive than equivalent types. For instance, in aerospace communications, literally dozens of transistors are used in parallel in some applications to obtain very high output.

In spite of this, transistors are not replacing vacuum tubes in all applications. The amateur operator who wants to put out 2,000 watts is not likely to use transistors except in the driver stages where the transistors can make a very compact and efficient assembly.

At the present time, low-power transistors are quite reasonable. Foı higher power a few devices are now becoming available on the surplus market. Most of them are not modern transistors, and suffer from many of the faults that I mentioned before, particularly failure due to mismatching or detuning. Nevertheless, they are quite useful in many applications and are a very good way to get your feet wet in rf power before you take on a more expensive project.

For that matter, dedicated hams have never had any real problems in obtaining components for their projects. The serious ham who wants to build a high power transistor transmitter can likely get the transistors he needs one way or another, just as he has been able to obtain expensive varactors for microwave use. And even though the transistors are relatively expensive, they are quite reasonable when you consider their advantages; using transistors that operate directly from a car battery, for example, eliminates the need for a relatively expensive, space-consuming inverter.

**fig. 2. The geometry of a Motorola balanced-emitter (resistor stabilized) transistor, the 2N5637, which is capable of 20 watts output (minimum) at 400 MHz. The 2N5637 is composed of 220 individual small transistors connected in parrallel, each emitter. This construction provides excellent safe area and resistance to damage from detuning or high vswr.**

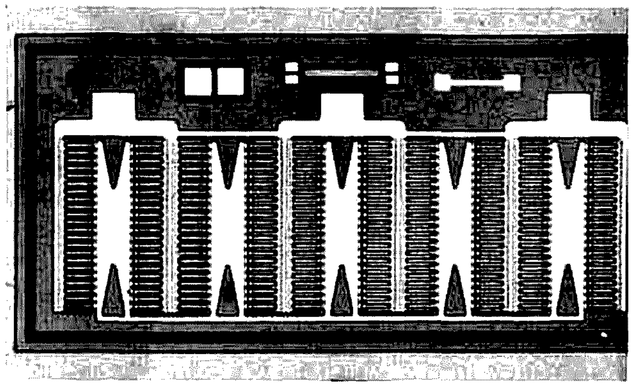
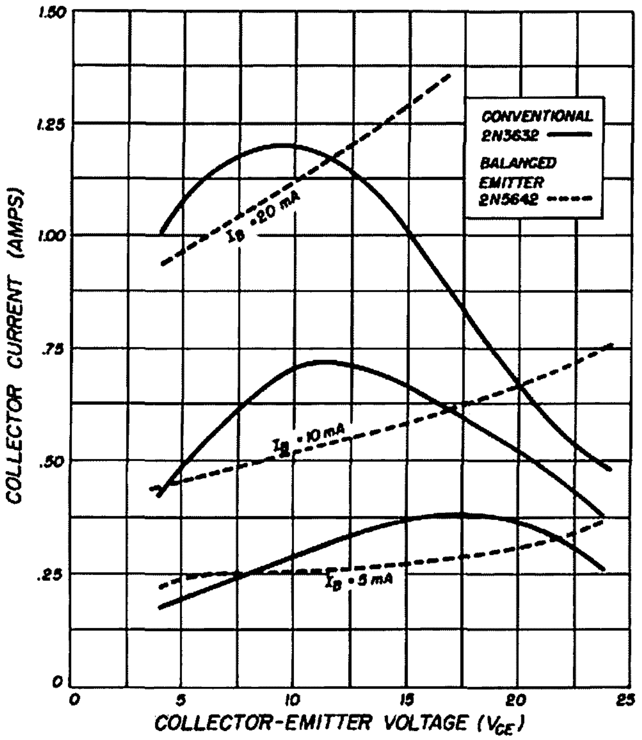


**fig. 1. Comparison of collector current vs collector-emitter voltage in conventional and resistor-stabilized transistors (balanced-emitter transistors).**



The principles outlined in this article apply equally well to small transistors used in 1- and 2-watt transmitters and to the large transistors that are necessary to get 100 watts or more of rf power output. The same design principles are used in all of these applications. The numbers will change, of course, and sometimes the networks used for coupling between the transistors will also change due to the differences in impedance levels. However, if you learn how to design a low-power transmitter you can apply the same principles when higher-power transistors become available to you.

## characteristics of rf power transistors

Modern rf power transistors are made of many individual small transistors in parallel. These transistors are formed at the same time in the manufacturing process. The small transistors are then connected in parallel with aluminum metal that is deposited on the surface of the silicon chip. Each of the small transistors handles relatively little power, hence, can be rather small in size. This is an advantage in high-frequency use.

A further development of this type of construction is the balanced-emitter transistor. Here a small resistor is placed in series with the emitters of the small transistors that are connected in parallel to form the whole transistor.

**Fig. 2** shows a typical balanced-emitter transistor. It is the Motorola 2N5637, which can supply an output of 20 watts at 450 MHz. This transistor consists of 220 transistors in parallel, and is stabilized by 220 small thin-film Nichrom resistors. This device, which is more complex than most ICs, is 50 by 100 mils (0.05 by 0.1 inches) in size. You'll notice that the 2N5637 is made of ten cells. Similar cells are used in other transistors: the 2N5636, which is often used as the driver for the 2N5637, consists of six cells and can provide 7.5 watts. The 2N5635 contains two cells and can put out about 2.5 watts.

The reason for this complex construction is that it improves ruggedness. If one small transistor in the large chip starts drawing more current than another one because of some small difference in its construction, the current through it would increase. Then the voltage across the small resistor would increase, increasing the emitter-base voltage. This reduces the amount of current that this individual transistor draws. In other words, it is a self-stabilizing operation. No single transistor can draw an excessive amount of current. This protects the transistor from secondary breakdown and permits it to stabilize itself in the event of severe load mismatch or circuit detuning.

Since these small emitter resistors are in parallel, their equivalent resistance is very small and does not result in significant degeneration or loss of gain. On the other hand, if a conventional, older type of power transistor is used with emitter-resistor pro-

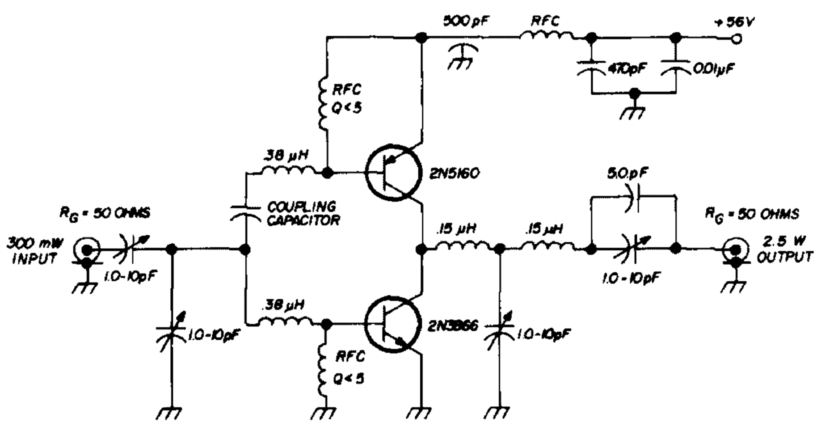



**fig. 3. 300-MHz complementary rf power amplifier using npn 2N3866 and pnp 2N5160 transistors.**

**table 1. Typical rf power transistors**

| Type | Supply voltage (cw service) | Gain (min dB) | $P_{out}$ (min W) | @ f (MHz) | Case | Single quantity cost |
|---|---|---|---|---|---|---|
| 2N3866 | 28 | 10 | 1 | 400 | TO-39 | $ 2.25 |
| 2N3375 | 28 | 8.6 | 7.5 | 100 | TO-60 | 10.80 |
| 2N3553 | 28 | 10 | 2.5 | 175 | TO-39 | 4.37 |
| 2N3632 | 28 | 5.9 | 13.5 | 175 | TO-60 | 12.75 |
| 2N4072 | 13.6 | 10 | 1/4 | 175 | TO-18 | 2.25 |
| 2N4073 | 13.6 | 10 | 1/2 | 175 | TO-5 | 2.70 |
| 2N4427 | 12 | 10 | 1 | 175 | TO-39 | 2.15 |
| 2N5160* | 28 | 8 | 1 | 400 | TO-39 | 6.75 |
| 2N5161* | 28 | 8.75 | 7.5 | 175 | TO-60 | 18.75 |
| 2N5162*+ | 28 | 6 | 30 | 175 | TO-60 | 27.00 |
| 2N5635 + | 28 | 6.2 | 2.5 | 400 | 144B | 7.50 |
| 2N5636 + | 28 | 5.7 | 7.5 | 400 | 144B | 22.80 |
| 2N5637 + | 28 | 4.6 | 2 | 400 | 145A | 57.50 |
| 2N5641 + | 28 | 8.4 | 7 | 175 | 144B | 6.40 |
| 2N5642 + | 28 | 8.2 | 20 | 175 | 145A | 21.30 |
| 2N5543 + | 28 | 7.6 | 40 | 175 | 145A | 40.40 |
| 2N5644 + | 12.5 | 7 | 1 | 470 | 145A-01 | 11.80 |
| 2N5645 + | 12.5 | 6 | 4 | 470 | 145A | 15.50 |
| 2N5646 + | 12.5 | 4.7 | 12 | 470 | 145A | 29.20 |
| 2N5589 + | 13.6 | 8.2 | 3 | 175 | 144B | 6.10 |
| 2N5590 + | 13.6 | 5.2 | 10 | 175 | 145A | 14.40 |
| 2N5591 + | 13.6 | 4.4 | 25 | 176 | 145A | 25.20 |
| MM1552 + | 27 | 7.8 | 75 | 150 | 145C | 67.50 |
| MM4018*+ | 12.5 | 10 | 1/2 | 175 | TO-39 | 2.20 |
| MM4019*+ | 28 | 10 | 2.5 | 175 | TO-39 | 6.50 |
| MM4020*+ | 12.5 | 11.5 | 3.5 | 175 | 208-1 | 8.05 |
| MM4021*+ | 12.5 | 7.0 | 15 | 175 | 208-1 | 19.50 |
| MM4022*+ | 12.5 | 5.5 | 25 | 175 | 208-1 | 30.00 |
| MM4023*+ | 12.5 | 5.4 | 40 | 175 | 208-1 | 49.40 |

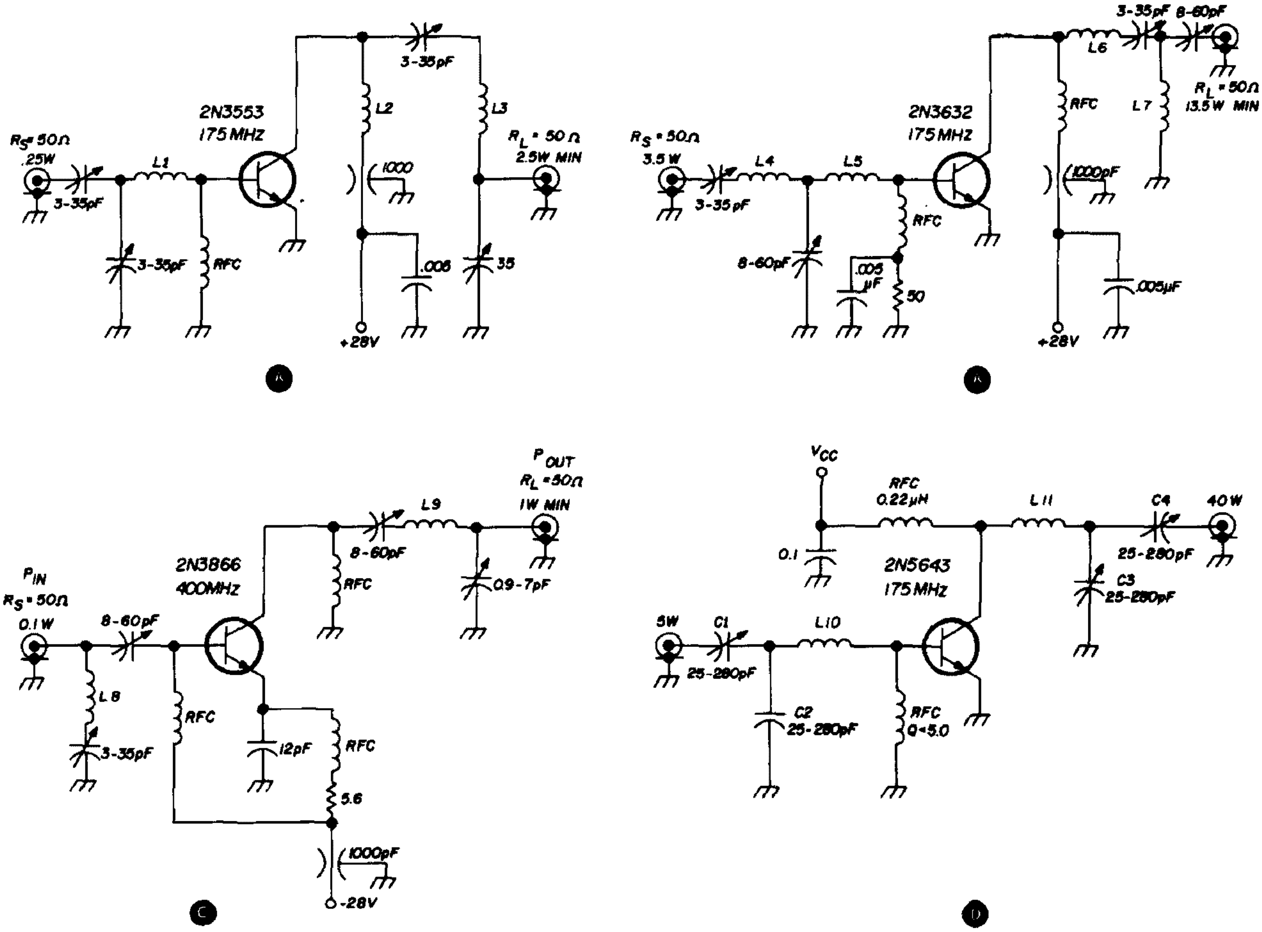
*pnp
+balanced-emitter transistor

tection, a resistor large enough to have any significant effect on the ruggedness of the transistor circuit would cause considerable loss of gain and output.

The greatest advantage of balanced-emitter transistors is their ruggedness. A balanced-emitter transistor can stand an infinite vswr for a short time in a-m service, for example. You can also tune one of these transistors without having it blow out, as often happens with older transistors.

Another result of this construction is shown in the $I_C/V_{CE}$ curve shown in **fig. 1.** Here the collector currents of two transistors with similar output capability are compared. One is a balanced-emitter transistor, the 2N5642. The other is a more conventional transistor, the 2N3632. The 2N3632 contains two chips in parallel in one package, but no emitter stabilizing resistors are included in this transistor. You'll notice that as the voltage increases in the balanced-emitter transistor, the current increases proportionately. This shows the excellent linearity which would make it ideal for amplitude modulation or linear amplification. The 2N3632 has a negative resistance region where increasing the voltage results in lower current. This negative resistance region would result in very poor upward modulation, of course, and high distortion in amplifier service.

While most silicon transistors, particularly power transistors, are npn devices, pnp rf power transistors are also made by Motorola. One, the MM4023, is a balanced-emitter transistor capable of 40 watts output at 175 MHz. The lower-power 2N5160 is a close pnp match of the popular 2N3866 and can be used in complementary service (**fig. 3**).




| | | | |
|---|---|---|---|
| **L1, L2** | **2 turns no. 16 AWG, 3/16" diameter, ¼" long** | **L8** | **1" straight no. 14 wire** |
| **L3** | **3 turns no. 16 AWG, 3/8" diameter, 3/8" long** | **L9** | **1 turn no. 16, 1/4" diameter** |
| **L4, L6** | **4 turns no. 18 AWG, ¼" diameter, 3/16" long** | **L10** | **2 turns no. 18, 1/4" diameter, 1/8"long** |
| **L5** | **1 turn no. 16 AWG, ¼" diameter, 3/16" long** | **L11** | **2-3/4 turns no. 18, 1/4" diameter, 3/16" long** |
| **L7** | **2½ turns no. 16 AWG, ¼" diameter, ¼" long** | | |

**fig. 4. Test circuits used for typical rf power transistors.**

**Table 1** summarizes a number of rf power transistors, both conventional and balanced-emitter types. The conventional ones are suitable for low power stages, for drivers, and where they will not be subjected to load mismatch or detuning. Some of these transistors are also becoming available at relatively low prices in surplus. However, only balanced-emitter transistors are recommended for use where they will be modulated, where any significant power is being handled, or for feeding an antenna. **Fig. 4** gives test circuits for some of the transistors. Many of these circuits can be adapted for use in the ham bands.

## types of operation

Hams are interested in rf power transistors for four modes of operation: cw, fm, a-m and ssb. The simplest of all of these is cw operation. Keyed cw can be used in portable operation where the maximum range is desired. There's no question that this operation provides you the best range for a given power. A continuous signal can also be used for driving varactor multipliers or vacuum tubes. Fm operation is the same as cw as far as a transistor is concerned. The deviation used in any type of ham or commercial communications work is so

small that it appears as a constant signal to the transistor.

In either cw or fm operation the transistor can be operated at a supply voltage of slightly less than the collector-emitter breakdown voltage ($BV_{CEO}$). For example, the 2N5641 series of transistors has a minimum $BV_{CEO}$ of 35 volts and it is quite suitable for use at 28 volts for fm or cw operation. Likewise, transistors with an 18-volt $BV_{CEO}$ can be used with the automobile supply, which is roughly 13.5 volts. Because you can operate relatively close to the breakdown voltage, you can get maximum power output from a transistor in cw or fm operation.

Incidentally, the collector voltage of a transistor rises to roughly twice the supply voltage during the cycle. This would seem to exceed the transistor ratings, but this is not true because the radio-frequency breakdown voltage is considerably higher than the dc voltage breakdown. It is very close to the highest maximum rating normally given on a transistor data sheet, the $BV_{CES}$.* The $BV_{CES}$ is 65 volts for the 2N5641 series.

Operation of a transistor at 28 volts requires an inverter if it is used in a car. This inverter can be relatively simple—even an autotransformer that provides voltage doubling. However, this partly negates one of the great advantages of using transistors: the fact that they can be operated directly from the 13.6 V supply voltage. These 28-V transistors are quite useful in fixed-station operation, but they are more often used in a-m service. A transistor operated at its maximum cw output, say 40 watts for the 2N5643, must be given some type of protection in case of extended detuning or mismatch. The transistor can survive a short fault but not a continuous one.

Transistors are available for operation from a car battery of 13.5 volts. They are quite similar to the higher-voltage devices but are optimized for maximum output at the lower voltage, and have lower breakdown voltages. They also have lower gain at the lower voltages. For example, the 2N5591 has an output of 25 watts at 175 MHz when operated directly from a 13.5 V supply. Its power gain at this level is only 4.4 dB minimum, which is relatively low. The 2N5642, which has roughly the same output, 20 watts at 175 MHz, has a gain of 8.2 dB when it's operated at 28 V. Because of this lower gain, more stages are generally required for the same power level with low-voltage power supplies.

*The $BV_{CES}$ is usually numerically about equal to the $BV_{CBO}$.

## amplitude modulation

Amplitude modulation with transistors is usually a rather messy proposition. Frequency modulation is much more satisfactory, and hams are using fm more and more in vhf mobile communications. However, a-m is widely used commercially in aircraft transmitters and by the military. The aircraft transmitters operate between 108 and 136 MHz, and the military use a-m between 108 and 152, and between 225 and 400 MHz. For this reason, many transistors have been developed for a-m use in these frequency ranges. The carrier output of a transistor in a-m service is very low compared to its cw output. For example, the 2N5643 can put out 40 watts on cw or fm at 175 MHz, but it's only suitable for about 15 W of a-m carrier. However, on the modulation peaks, this increases to about 60 watts PEP, of course.

In a-m operation you have to operate a transistor at less than half its collector-emitter breakdown voltage. For example, the 2N5643, which can be used at 28 V for cw operation, cannot be operated at more than about 14 V in a-m service; this is because on a-m peaks the voltage rises to twice the normal maximum, which is already twice the supply voltage. In other words, on a-m peaks a 13.5 V supply will give rf peaks that rise to 54 V. A transistor that is to be used in a-m service at 13.5 volts, then, must have a $BV_{CES}$ greater than 54 V.

As you can see, an amplitude-modulated transistor has to be operated at about one-half its normal supply voltage, where it provides maximum gain. Its gain will be lower than that of a transistor made specifically for 13.5 V service. Amplitude modulation involves a number of compromises; it is used only because a-m equipment is already

very popular and widely used. Fm is far more satisfactory with transistors; it also provides much greater range for the same power inputs.

It might be noted, however, that large aircraft which use a-m are using transistors—single transistors such as the MM1552 which is suitable for 25 watts carrier output at 135 MHz with 100 watts peak power. The MM1552 is capable of about 75 W carrier output in cw operation. This particular transistor is used in a-m service at 13.5 V, and has a breakdown voltage of about 65 volts.

In a-m service, because the transistor is operated at relatively low carrier output, it can withstand infinite vswr and detuning for a considerable period of time if mounted on an adequate heat sink.

## ssb

Single sideband with transistors is still relatively unfamiliar to most users. Transistors have been used for single sideband for some time, particularly by the military, but not too much information is available on this type of operation. A rule of thumb is that a transistor provides fairly low distortion at a peak envelope power output roughly equal to the cw rms output. As an example, the 2N5643, which can put out 40 watts of cw, can provide 40 watts PEP of sideband with relatively low distortion.

Balanced-emitter transistors are ideal for single sideband because of their excellent linearity. At the present time an inexpensive transistor can provide about 8 to 10 watts PEP ssb, making it quite suitable for use alone or to drive an efficient transmitting tetrode tube such as the 4CX1000. This is not enough output, of course, to drive a grounded-grid tube like the popular 3-1000Z.

**Table 2** summarizes the required voltage ratings of transistors used at 13.5 V and 28 V in all popular modes.

## reading data sheets

An important part of using rf power transistors is understanding their data sheets. Data sheets on any power transistor or for that matter, any semiconductor, are available from the manufacturer of the device.* Most of the data sheet is quite straightforward, and though different manufacturers use different formats, similar information is available from most data sheets. One of the first things that you should remember when you are looking at a data sheet is that there are different types of values given. Some are actual maximum ratings. These are the absolute limits to which a transistor should be subjected. Other values are characteristics which describe the actual performance of the transistor.

In the maximum ratings there is no problem about interpreting them; they are quite obvious. However, the characteristics can be **typical** values, or they can be **minimum** or **maximum** values. The manufacturer chooses the value to give him a reasonable yield of salable devices. At the same time, most of the transistors that he produces exceed the minimum ratings, sometimes by quite a bit. For this reason, typical values are often given on data sheets. These typicals include all of the curves, except one or two such as the safe operating area curve and temperature deratings.

Typical values are very useful in design; however, it's better to design with the minimum values to be on the safe side and insure that your design works properly. The data sheet clearly differentiates between typical and minimum values.

Among the curves which provide typical values are those giving impedances, where it is not practical to give a range. In this case, many transistors are measured, and an average value is put on the curves. These values can vary a bit in individual transitors, but the numbers indicated are usually quite close and satisfactory for circuit design.

One of the first ratings or characteristics that you are concerned about is the breakdown voltage of the transistor as discussed in the section on classes of operation. Many different breakdown voltages are provided on data sheets. The most significant one for rf use is the $BV_{CES}$. If this is not provided, the $BV_{CBO}$ is usually numerically about the

*Data sheets on any transistors mentioned in the text are available from Technical Information Center, Motorola Semiconductor Products Inc., Box 20924, Phoenix, Arizona 85036.

same. Half of this value gives you the maximum rating for cw or fm use; one-quarter of it for a-m use, as shown in **table 2.**

It's interesting to notice the trade-offs that accompany a higher breakdown voltage in a given family of transistors. A higher breakdown voltage indicates a lower output capacitance or $C_{OB}$. This, of course, can simplify design at high frequencies considerably by reducing the amount of parallel output capacitance. An unfortunate result of higher breakdown voltage is higher dc and rf saturation voltages. The reason this is important is that the actual output from a transistor is dependent on the collector voltage swing, or difference between the collector supply voltage and the saturation voltage.

**table 2. Minimum $BV_{CEO}$ and $BV_{CES}$ for translators used in various modes of operation at 13.5 and 28 V. Values for a-m assume 100% modulation.**

| | 13.5 V Supply | | 28 V Supply | |
|---|---|---|---|---|
| | $BV_{CES}$ | $BV_{CEO}$ | $BV_{CES}$ | $BV_{CEO}$ |
| cw | 30 | 15 | 60 | 30 |
| fm | 30 | 15 | 60 | 30 |
| a-m (transformer modulation) | 60 | 30 | 120 | 60 |
| a-m (series modulation) | 30 | 15 | 60 | 30 |
| ssb (linear amplification) | 30 | 15 | 60 | 30 |

For example, though dc saturation voltages are rarely given, for rf power devices they typically run around 1.5 to 2 volts for high voltage (28 V) transistors, and a little bit lower for low voltage ones. However, the rf saturation voltage is usually about 1.3 times higher and this reduces your power output. As you can see, if you operate a transistor with a high breakdown voltage at a low voltage, you reduce your voltage swing considerably because the high rf saturation voltage will remain roughly the same. Thus, a high breakdown voltage results in a lower maximum saturated power output. But as discussed before, a high breakdown voltage is a necessity for amplitude modulation, and so we have to live with the high saturation voltage that accompanies it. This is another good reason to use fm rather than a-m.

Incidentally, at high operating voltages, gain is higher than at lower voltages, partly because the higher operating voltage reduces both output and feedback capacitance.

One parameter that is of relatively little importance is the maximum collector current ($I_{C(max)}$). Though a safe operating area graph often lists the maximum permissible simultaneous voltage and current for the transistor, these values are usually dc or low-frequency ones and are not very relevant at 100 MHz or so. Transistors aren't often operated near their maximum collector currents, anyway, whether they are low-frequency or high-frequency devices.

A vital parameter in a high-power amplifier is the maximum power dissipation. The maximum power dissipation of a transistor is the difference between the input and the output: $P_D = P_{in(rf)} + P_{in(dc)} - P_{out}$. For example, if you have 1 watt of rf input and 10 watts of dc input (a total of 11 watts input) and five watts output, the dissipation is 11 minus five, or six watts. If you're using a relatively large transistor it may be able to handle this with very little extra heat sinking; however, it is important that sufficient heat sink be provided if necessary.

Dc current gain or $h_{FE}$, is relatively important in many applications, but its significance in rf power transistors is probably not what you think. A high $h_{FE}$ indicates a high $f_T$ and hence a high power gain at frequencies below the $f_T$. Nevertheless high $h_{FE}$ is not desirable in most rf power transistors: it results in lower maximum saturated power output, higher intermodulation distortion in single sideband use, greater change in dc gain with changing current and, perhaps most important, dc and low-frequency instability.

The lower dc stability means that it is relatively hard to stabilize the bias of the transitor in class B or AB operation for ssb. The ac instability can lead to low-frequency oscillation because the transistor has so much gain at these frequencies in comparison with the gain at the very high frequencies at which you want it to operate.

It follows that a high $f_T$ is not necessarily

an advantage. The $h_{fe}$ (small-signal ac current gain) and $f_T$ are intimately related, since $f_T$ is equal to $h_{fe}$ times the frequency at which $h_{fe}$ is measured. High $f_T$ means higher output resistance in a transistor. Higher resistance can simplify matching requirements in some cases but the high $f_T$ also means a lower input resistance at a given frequency and a lower maximum saturated power output. *All* in all, $f_T$ is not really a very good indication of a transistor's performance in power amplifying service.

The important numbers for you to look for in an rf power transistor are its functional tests. Rf power transistors undergo tests for gain, power output, and in some cases, efficiency, at given frequencies. This is a rather time-consuming, and hence, expensive, operation for the manufacturer and one of the reasons that rf power transistors are more expensive than low frequency ones. However, it insures that the transistors are suitable for high-frequency operation.

The functional test can be given in a number of different ways; probably the most obvious one is a minimum power output for a given power input at a given frequency. A more common test furnishes the amount of input required for a given output. Power gains are usually given at the same frequency at which the power outputs are measured. Minimum and typical values are often given. The minimum is what you should design with; the typical is what you can hope for.

If you do a little bit of figuring, you will find that most power transistors have much lower gain than vacuum tubes you are familiar with. Therefore, more transistors than tubes are required to obtain a given power level in most cases. This is not necessarily true at relatively low frequencies: a power transistor can have very high gain at 50 MHz, for example, if it's designed for use at 400 MHz. Power gain increases about 6 dB per octave, and this can mean that you have much higher gain at lower frequencies.

However, it's not necessarily desirable to use a 400-MHz transistor at 50 MHz. If you have excessive gain you're likely to have instability. In general, about 15 dB is the maximum gain you should expect to get out of an rf power transistor and have it remain stable. More than this and you're likely to be bothered by instability that could be hard to eliminate. In general, you should use rf power transistors only in the ranges that are indicated on the data sheet. For example, if output powers and impedances are given for a transistor between 100 MHz and 400 MHz, you could use it anywhere within that range and probably just a little bit above or below it. However, it would be best not to use this transistor at 30 MHz or below.

A relatively recent development in rf power transistor data sheets is the inclusion of large-signal impedances. Previous to this only small-signal impedances were given: a 20-W transistor might be characterized in a circuit in which it was actually just a low-level amplifier. However, when transistors are operated at high power levels, their characteristics are quite different from those at low power levels.

**Table 3** lists the high- and low-level impedances for the 2N3948 transistor at 300 MHz. You can see the vast differences between these values. If you use the small-signal impedances to design a transmitter, it won't work properly. Some manufacturers still do not give large-signal impedances, complicating the task of the designer considerably, because he must spend a great deal of time in empirical work. Incidentally, Motorola pioneered in providing large-signal impedances, and they are provided on almost all Motorola rf power transistor data sheets.

Three different large-signal impedances are provided: the input capacitance ($C_{in}$), input resistance ($R_{in}$), and output capacitance ($C_{out}$). The output resistance ($R_{out}$) can be figured from the supply voltage and output power of the specific circuit you are using, and that will be discussed in more detail further along. Incidentally, the output capacitance is roughly twice the low frequency $C_{OB}$ in case this is not given.

There are two different ways that impedance data can be presented: in the parallel form, which is given on most Motorola data sheets, or in the series representation. A parallel form would be, for example, 6 ohms resistance in parallel with 30 pF capacitance.

The series form would be the familiar expression using j, such as 25 — j8 ohms. There are advantages to using either form; some networks are easier to design with the series representation, and some with parallel. It is relatively easy to switch from one to another. Later on in the discussion of network design I will indicate when you use the series and when you use the parallel form, and how you change from one to the other.

## packaging

The package for an rf power transistor is vitally important. For large power outputs, specialized packages that provide minimum lead inductance are required. Though the TO-39 package is widely used for low-power transistors such as the 2N3866 and the 2N3553, it is not suitable for powers over a few watts. The next step up is similar to the TO-39 except it provides solid terminals instead of wire leads and uses a stud for mounting (TO-60). Examples are the 2N3375 and 2N3632. These packages are shown in **fig. 5**.

A much-better package is the strip-line opposed-emitter case, which is used in one form or another by most manufacturers. This type of package provides an isolated stud for mounting. This stud may be mounted directly on a heat sink without insulating washers. Four ribbon leads are provided; two emitter leads, a collector lead and a base lead. The two emitter leads are between the collector and the base leads providing excellent isolation, and the fact that there are two of them makes it easy to provide a very low impedence ground. A wide ribbon is used for high power levels and a smaller one for lower power levels. The Motorola strip-line package is ceramic; some of the others are plastic. The most popular package is only 3/8 inch in diameter, yet can put out over 40 watts of power.

## circuit design

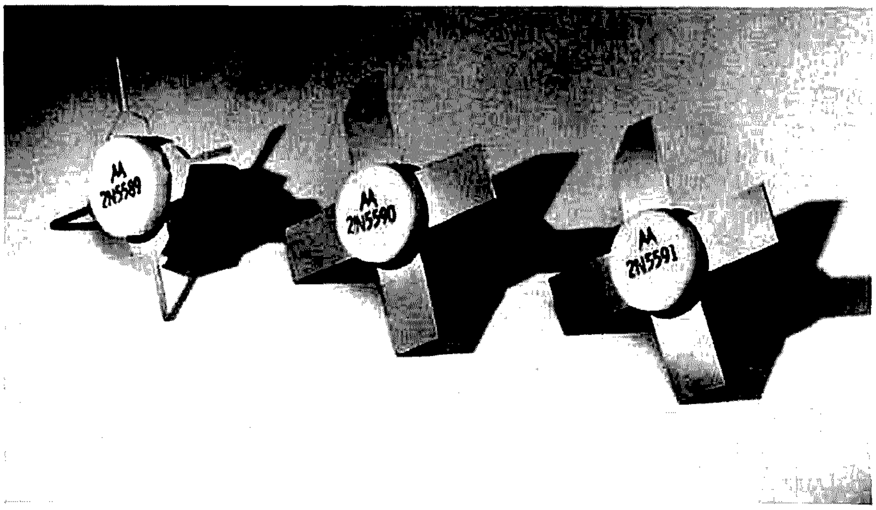
Hams are fortunate in at least one respect when it comes to rf power transistors: most amateur circuits are narrow band, unlike the wideband transmitters required in commercial and military a-m service. In broadband circuits, considerable gain often has to be sacrificed to obtain the wide band. However, hams can use the transistors in narrowband service and obtain the performance specified on data sheets without any great problem.

The first problem that a transmitter designer must solve is the frequency at which he will generate his signal, and at what level any frequency multiplication, if that is needed, will be performed. General commercial practice seems to obtain a low-level signal at the output frequency, then perform all the power amplification at this frequency. There are a number of reasons for this: one is that in many commercial applications a frequency synthesizer is used, and its output can conveniently be at the output frequency. Another reason is that it is easier to design an amplifier stage than a multiplier. The information required for designing an amplifier can be obtained very readily from a data sheet, while that for designing a multiplier often must be obtained by cut and try. For this reason, it's usually best to plan on having a few milliwatts, say 20 to 100, at the output frequency and amplifying from there. All of the multiplication that's needed can then be done at a low level.

The next problem is whether to use a low-frequency crystal and multiply up, or to use a higher frequency crystal. A low-frequency crystal is usually necessary in fm applications where you need to use a relatively low frequency and multiply by a fairly high number to get enough deviation for fm. However, for a-m or cw, it's usually best to use as high a frequency as is practical. Since very little power output is needed, you can use an overtone crystal and just multiply a few times. For instance, for two-meter output a 72-MHz overtone crystal oscillator can provide a few milliwatts which then can be doubled. This is usually the simplest approach; more important, this high-frequency signal generation reduces the number of harmonics and sub-harmonics that you have to contend with. It's relatively difficult to eliminate frequencies every 8 MHz across the band, but easy to suppress ones that are 72 MHz from the desired frequency.

This discussion, of course, has been as-

**New Motorola balanced-emitter transistors in a ceramic stripline package provide up to 20 watts output at 400 MHz (2N5637), 40 watts at 175 MHz with a 28-volt supply (2N5643), or 25 watts at 175 MHz with a 13.5-volt supply (2N5591). Also available are new transistors that are suitable as drivers for those devices or as lower-power amplifiers.**



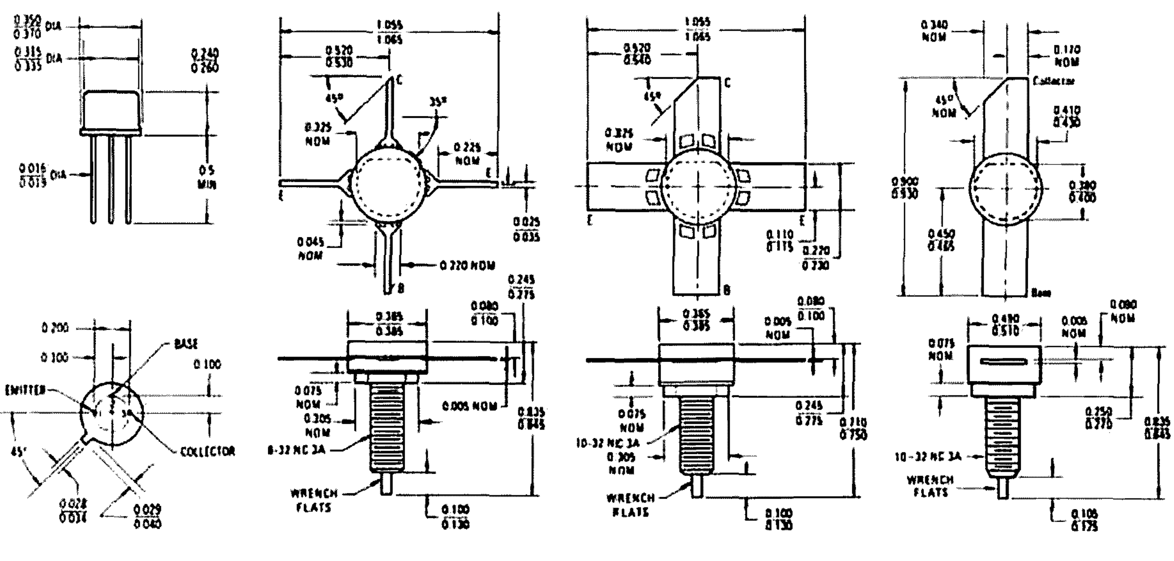

suming that you are using crystal control. If you use a variable-frequency oscillator, you'll have some other problems. Then your best bet is to use the heterodyne method so that your vfo operates at a relatively low frequency and beats against a relatively high-frequency crystal oscillator. For single sideband, of course, this is a necessity.

## transistor selection

Choosing transistors for use in a transmitter can be an interesting task. In many cases, you really have very little choice. You may have a few transistors of a given type, or you may be limited in the amount you can spend for transistors. In this case, your choice will be relatively limited. And considerably simplified, for that matter. In other cases, you'll have to decide the power output you want, taking into account the power supply that is available, and work backwards from this. As a practical example, a simple transmitter for two meters will be developed in the rest of this article. This transmitter will also be used to explain simple network design.

Suppose we would like to obtain about 10 watts of cw or fm on two meters to drive a fixed station amplifier. A 28-volt supply will provide the highest output. A good transistor choice would be the 2N5641. It has a minimum power output of 7 watts at 175 MHz according to **table 1.** Referring to the data sheet, it can be seen that its output at 145 MHz would be much closer to 10 W. This transistor costs $6.40 in single quantity,

**fig. 5. Typical packages for rf power transistors; from left to right: TO-39, case 144B, case 145A, and case 145C. The 145C case is a ½" case; the others are ⅜".**



**tabia 3. Small- and large-signal performance data for tha 2N3948 at 300 MHz show tha inadequacy of using small-signal characterization data for large-signal amplifier design. Resistances and reactances shown ara parallel components. That is, the large-signal input impedance is 38 ohms in parallel with 21 pF, etc.**

| | class A Small-signal amplifier $V_{CE}$ = 15 Vdc; $I_c$ = 80 mA | class C Power amplifier $V_{CE}$ = 13.6 Vdc; $P_o$ = 1W |
|---|---|---|
| Input resistance | 9 ohms | 38 ohms |
| Input capacitance or inductance | 0.012 μF | 21 pF |
| Transistor output resistance | 199 ohms | 92 ohms |
| Output capacitance | 4.6 pF | 5.0 pF |
| Power gain | 12.4 dB | 8.2 dB |

a reasonable price for a transistor of this output. Table 4 summarizes the most important characteristics of this transistor at 145 MHz; the values were simply taken from the appropriate graphs on the data sheet.

At this frequency, the 2N5641 has an output of 9 watts for an input of 0.5 watt. To be on the safe side, we can use the 2N3866 as a driver. It has an output of 1 watt at 145 MHz with only 20 milliwatts of input, a gain of about 17 dB. This high gain is safe in this low-level stage, and should not cause any problems. A block diagram of the transmitter is shown in **fig. 6.**

The 20 mW of drive can be supplied by a small-signal transistor, such as a plastic-encapsulated MPS3563, an excellent transistor for this use, costing only $0.44.

For high power levels, paralleled transistors might be needed. If this is done, some type of equalizing network must be provided to insure that both transistors receive the same drive. It's usually very difficult to use larger or better antenna or lower loss lead-in to get this gain in transmitted output.

The transistors discussed in this article generally operate in class C. In usual transistor practice, this means they are operating without any bias except that provided by the signal, without respect to the angle of conduction. Class-C amplifiers give excellent efficiency and high power output. They are also self-protecting: if you remove the drive from a class-C amplifier, it cuts itself off and does not draw current.

Slightly more gain can be obtained from class-A, -AB or -B amplifiers, but only at the expense of higher dissipation and smaller output. These other classes of operation can provide linear operation: hence they can be used for amplifying ssb or a-m. A class-C amplifier can be used only for amplifying cw or fm.

Normally a class-C amplifier has a choke or rf coil connected directly between the base and emitter (ground), but sometimes a

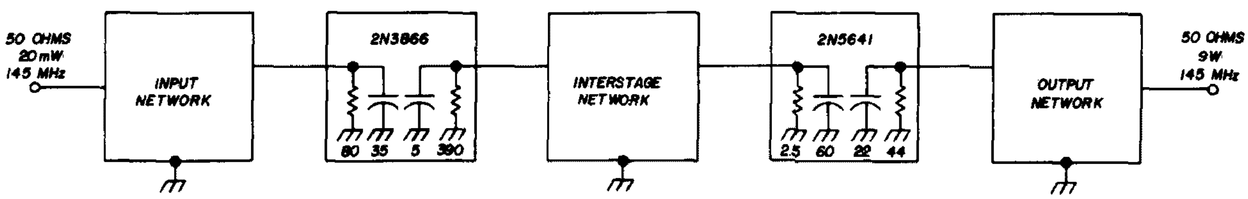



**fig. 6. Block diagram of a 9-watt transmitter for two meters.**

push-pull because of the problems in getting balanced drive. However, it should be remembered when considering this that only about 3 dB is gained by using another transistor in parallel. It might be easier to use a small resistor is connected in series with the base choke. This improves efficiency slightly at the expense of gain and output. This higher efficiency is normally not required except in battery operation or where

there might be problems with heat and the higher efficiency would reduce the power dissipation.

Transistor rf power amplifiers are usually not neutralized. Neutralization of a transistor is difficult because its capacitances vary greatly with applied voltages. Almost the only type of neutralization that is used is emitter tuning. Here a small capacitor is connected from the emitter to ground and tuned for maximum output. A small choke can be placed in the emitter lead, or, at the highest frequencies the emitter lead can provide sufficient inductance by itself.

This emitter tuning can provide higher output and higher power gain, but possibly at the expense of instability. Emitter tuning is a narrow-band technique and not suitable for most commercial use. Hams can use it because it is not too difficult to tune up one transistor for maximum power output. However, more conservative design does not accept emitter tuning.

Grounded-emitter operation is almost universal in rf power design. The grounded-base configuration is less stable, and adjustments for grounded-base amplifiers are more critical. If neutralization is required, it's very difficult to implement. Grounded-base amplification might be desirable in some applications, but grounded-emitter stages are usually much more satisfactory. In fact, transistors such as those in the strip-line opposed-emitter package have two emitter leads which are connected directly to ground. These transistors would not be very convenient for grounded-base operation.

In some rf power transistors, the emitter is internally grounded to the stud which helps reduce emitter inductance when the chassis is the rf ground. However, where the transistor is placed through a hole in a circuit board, the two emitter leads can provide shorter ground paths than an emitter connected to the stud.

**table 4. Characteristics of 2N3866 and 2N5641 at 145 MHz and 28 V.**

| Type | Input | Output | Gain | $R_{IN}$ | $C_{IN}$ | $C_{out}$ |
|---|---|---|---|---|---|---|
| 2N5641 | 0.5 W | 9 W | 12.5 dB | 2.5 ohms* | 60 pF* | 22 pF* |
| 2N3866 | 20 mW | 1 W | 17 dB | 80 ohms** | 35 pF** | 5 pF** |

*at 7 W output

**at 1 W output

## matching networks

Matching networks are used at the input and output of a power amplifier and between transistor stages. These matching networks serve two functions: impedance transformation and frequency selection. They provide an impedance transformation between the source and input, between the output load and load, and between stages. If a transistor had exactly 50 ohm input impedance or output impedance, the network could be very simple, simply a large capacitor. However, in practice the impedances are usually quite different from 50 ohms. In high-power transistors, the input impedance is often less than 1 ohm, and the output impedance only slightly larger.

The matching network also discriminates against unwanted frequencies. A simple network usually cannot provide sufficient discrimination, and it's always desirable to use an antenna filter with any type of transmitter that you connect to an antenna.

Transformer or loop coupling is rarely used in transistor rf power amplifiers. This type of coupling is hard to adjust for maximum power output and maximum power transfer, particularly at higher frequencies. Instead, simple T networks and L networks are commonly used. Pi networks are rarely used in transistor stages because they often result in impractical component values, such as 0.5 pF capacitance or 20 nH* inductance, whereas other networks give practical values that can be used in a transmitter.

Tuned lines and coaxial cavities provide high efficiencies and frequency discrimination, but they are very bulky at vhf and are

*The nanohenry, abbreviated nH, is one-thousandth of a microhenry, so 20 nH = 0.020 μH.

rarely used for this reason. In the uhf region, circuits are often built with strip-line techniques. These copper lines deposited on ceramic or high-frequency circuit board give excellent results and are used in many commercial and military applications.

## selecting Q

An important part of any rf network design is choosing the loaded Q. A loaded Q between 4 and 12 provides a good compromise between various considerations. It provides convenient values with most networks, sufficient harmonic attenuation, good efficiency and smooth tuning. The loaded Q, incidentally, is quite different from the unloaded Q of the components. The loaded Q is dependent on the reactance of the components and the output resistance of the transistor. On the other hand, the unloaded Q is determined by the Q of the coils or capacitors and is far higher.

The efficiency of a network depends on the ratio of unloaded Q to loaded Q. Low loaded Q provides easy tuning and high efficiency, but it also provides poor harmonic attenuation. Very high loaded Q's provide excellent attenuation of harmonics but result in critical tuning and high circulating currents which usually result in poor efficiency with practical coils and capacitors. Since an output filter must be considered a necessity in modern operation, the actual value of Q is not critical.

## network design

The next step is designing the required matching networks. There are a number of approaches to this problem. Perhaps the easiest is using an admittance chart but it is a little involved for this discussion. Another convenient one is the Motorola application note, "Matching Network Designs with Computer Solutions," by Frank Davis.[1] This application note is very easy to use; you simply figure out what kind of network you want to use, which is dependent largely on the values you have to match, and look up the proper values in a table.

I highly recommend that you get a copy of this note if you are going to be doing any transmitter designing. The note includes tables for designing with a number of different types of networks. However, this note is not necessary for circuit design; it can be solved with simple mathematics.

The most commonly used networks are shown in **table 5** with the formulas that are used for solving them. Some of these networks are shown with solutions for a 50-ohm load or source; others are suitable for matching any impedance to any other impedance within certain limitations. Be sure to take note of these limitations: some output networks are only suitable for matching impedances below 50 ohms to 50 ohms; others can be used only for impedances above 50 ohms; still others can be used for matching a wide range of values to 50 ohms.

A point to notice is that some of these networks call for a series representation of the transistor representation. The equations used for converting from series to parallel and from parallel to series are given in **table 6.**

Often in a solution of one of these networks the component values that are obtained are not very practical. If this happens another type of network will have to be chosen. In some cases it may be necessary to use two networks in series to obtain a practical *impedance* transformation.

You may have noticed in **table 4** that the values of the collector resistance for the two transistors were not given. These values are best computed from the power output of the stage and the supplv voltage:

$$R_L' = \frac{(V_{CC})^2}{2\ P_o}$$

where $R_L'$ is the output resistance of the transistor, $V_{CC}$ is the supply voltage, and $P_o$ is the power output.

This is an approximation and does not account for the rf saturation voltage, but it's accurate enough for design. With this formula it is easy to figure the output resistance of the two transistors: for the 2N3866, $R_L' = 28^2/(2x1) = 390$ ohms; for the 2N5641, $R_L' = 28^2/(2x9) = 44$ ohms.

The next step is to determine what types of network should be used to match the input to the driver transistors, the driver tran-

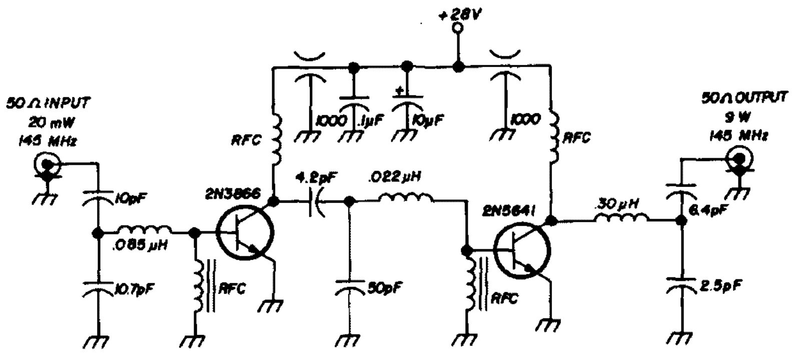



**fig. 7. This 9-watt transmitter for 145 MHz illustrates circuit design. In practice variable capacitors would be used, of course.**

sistor to the output transistor, and the output transistor to the load.

Referring to **table 5**, it appears that the most suitable network to match the output impedance of the 2N5641 to the 50-ohm load is the one shown in **table 5A.** The same network is also useful as an input network. Note that to compute this network the transistor output impedance should be in series form rather than the parallel form given on most of the data sheets and in **table 4.** Use the equations given in **table 6B** to convert from parallel to series representation. Incidentally, the reactances here can be figured most easily from a reactance rule such as the Shure rule, or from a table.

Now let's go through the whole design procedure using the steps listed in **table 5A:**

**1.** Convert the parallel form to series (see **table 6B**):

$$R_s = \frac{R_p}{1 + \left(\frac{R_p}{X_p}\right)^2}$$

To use this formula we need to find $X_p$, the reactance of a 22 pF capacitance at 145 MHz.

$$X_p = \frac{1}{2\pi fC} = \frac{1}{2\pi\,(145 \times 10^6)\,(22 \times 10^{-12})}$$
$$= 50 \text{ ohms}$$

This can also be found with a reactance slide rule or table.

Therefore,

$$R_s = \frac{44}{1 + \left(\frac{44}{50}\right)^2} = 25 \text{ ohms}$$

$$X_s = R_s\left(\frac{R_p}{X_p}\right) = 25\,\frac{44}{50} = 22 \text{ ohms}$$

**2.** Let $Q_L = 5$. This will provide adequate harmonic attenuation and practical component values.

**3.** With $R_o = 25$ ohms and $X_{Co} = 22$ ohms by step 1, calculate:

$$B = R_o\,(1 + Q_L^2) = 25\,(1 + 5^2) = 650$$

$$A = \sqrt{\frac{B}{R_L} - 1} = \sqrt{\frac{650}{50} - 1} = 3.5$$

**4.** Then

$$X_L = Q_L R_o + X_C$$

and L = 300 nH by a reactance chart or by $X/2\pi f$.

$= AR_L = (3.5)\ 50 = 175$ ohms

and $C_2 = 6.4$ pF (by a reactance chart or rule)

$$X_{C_1} = \frac{B}{Q_L - A} = \frac{650}{5 - 3.5} = 430 \text{ ohms}$$

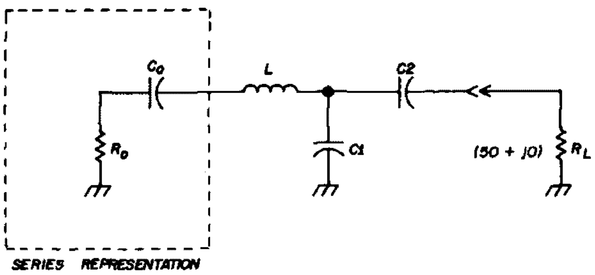
and $C_2 = 2.5$ pF.

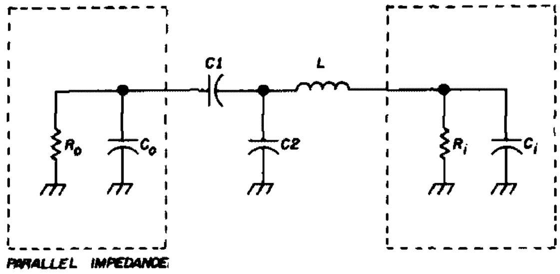
Similar computations are performed for the input and interstage networks. A Q of 5 is also useful here. The complete circuit of the transmitter is shown in **fig. 7.**

Once you have determined the proper inductance values for the transmitter coils you must obtain the coils. For low-frequency circuits commercially available inductors can often be used. However, for most vhf use you must wind your own. Most radio handbooks give instructions for this simple operation. Use large wire sizes for lowest losses and be sure to check the inductance with a dip meter and known capacitor.

Other transmitters designed with similar networks are shown in **fig. 8** and **fig. 9.** They illustrate the capabilities of modern rf power transistors.

**table 5. Matching networks.**




**A. Input or output matching network.**

1. **Convert the parallel form of impedance to series form if needed.**
2. **Select a $Q_L$ (usually 5 to 10; see text)**
3. **Compute:**

$$B = R_0 (1 + Q_L^2)$$

$$A = \sqrt{\frac{B}{R_L} - 1}$$

4. **Then**

$$X_{C2} = AR_L$$

$$X_{C1} = \frac{B}{Q_L - A}$$

**B. Interstage matching network. This network is useful when Ro is greater than Ri (which is almost always true).**

1. **Select a $Q_L$**
2. **Compute $A = Ri\,(1 + Q_L^2)$**
3. **Then $X_L = Q_L\,Ri$**

$$X_{C1} = X_{Co}\sqrt{\frac{A}{Ro} - 1}$$

$$X_{C_2} = \frac{A}{Q_L - \sqrt{\frac{AR_o}{X_{Co}}}}$$

## amplitude modulation

If you are building an a-m transmitter the modulation system is quite important. Low-level modulation is not recommended because it's inefficient. There are two major methods of high-level modulation of an a-m transmitter, transformer modulation and series modulation. Series modulation requires a supply voltage of twice the voltage required for the transmitter: an audio-frequency power transistor in series with the supply to the ouput stage of the transmitter operates as a variable resistance modulating the transistor output of the transmitter. This method does not use any transformers, but it requires twice the supply voltage that is needed for transformer coupling.

Transformer coupling is more conventional but it is usually difficult to find a suitable modulation transformer. Since relatively high current passes through the windings, a special transformer must be made in cases where the power levels over a watt or two.

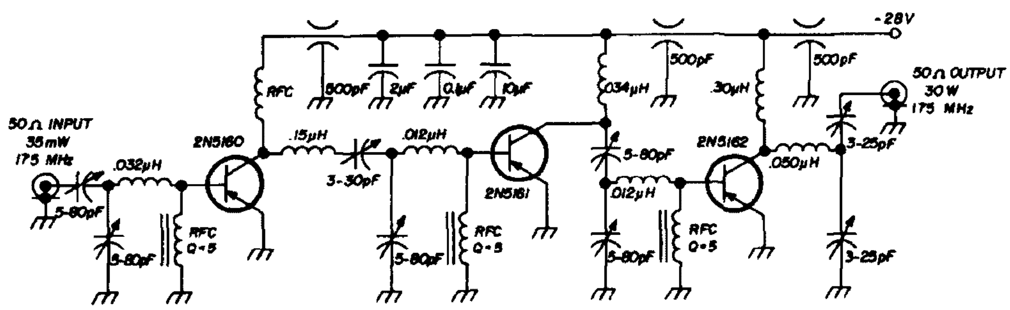



**fig. 8. 30-watt 175-MHz transmitter uses pnp transistors (from Motorola Application Note AN-481).**

You also have to be careful in transformer coupling so you don't apply too much supply voltage to the rf power transistor.

It is usually necessary to modulate not only the output stage in a transistor transmitter but also the driver, and in some cases previous stages. This can be done by applying full modulation to the output, partial modulation to the driver, and only upward modulation to the pre-driver, as shown in **fig. 10.** The diodes limit the modulation applied to the predriver stage to upward modulation.

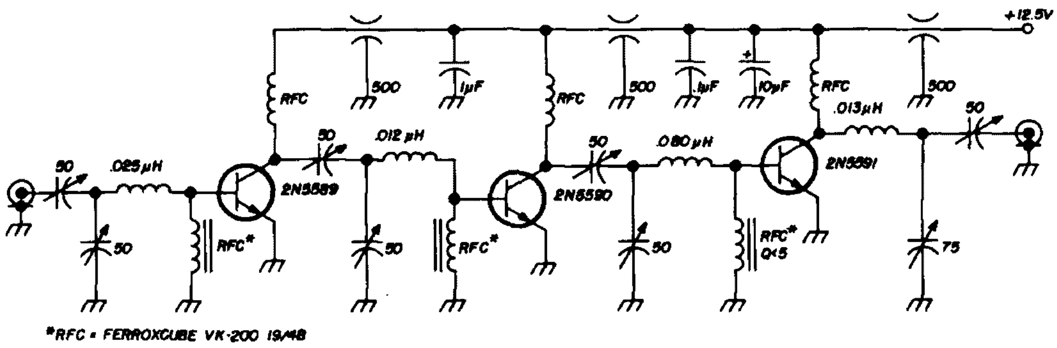
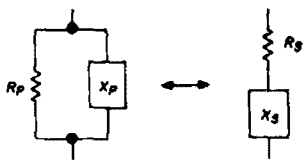



**fig. 9. 25-watt, 175-MHz transmitter designed for a 12.5-volt power supply (from Motorola AN-495).**

**table 6. Series-parallel conversion.**




**A. To convert a series representation of impedance to a parallel combination of resistance and reactance:**

$$R_P = R_S \left[1 + \left(\frac{X_S}{R_S}\right)^2\right]$$

$$X_P = \frac{R_P}{X_S/R_S}$$

**B. To convert a parallel combination to its series equivalent:**

$$R_S = \frac{R_P}{1 + \left(\frac{R_P}{X_P}\right)^2}$$

$$X_S = R_S \frac{R_P}{X_P}$$

**where $R_P$ is the parallel resistance, $R_S$ is the series resistance, $X_S$ is the series reactance, $X_P$ is the parallel reactance.**

**$X = 2\pi fl$. for inductance**

**$X = \frac{1}{2\pi fC}$ for capacitance**

Modulating all these stages is necessary because the gain of a power transistor is low enough that there is significant feedthrough from earlier stages. For example a transistor with 10 watts of ouput may have another watt contributed by the driver stage. If this stage is not modulated it will limit the maximum possible percentage of modulation.

## thermal design

An important part of the design of high-power transistor transmitters is its thermal aspects, or determining what size heat sink should be used to prevent the device from getting too hot and destroying itself. For relatively low-power transmitters this is not a great problem, and connecting the stud to a metal chassis is adequate for powers below about 15 to 20 watts. For higher powered transmitters, more attention should be paid to this topic. Thermal design at rf is similar to that at lower frequencies. However, the heat sink must also provide a good path for rf in some types of construction. Provision may also have to be made to dissipate considerable extra heat during periods of mismatch or detuning.

## *practical construction*

An important part of building a transistor transmitter, particularly for the vhf range, is using very short leads. The fact that wide

ribbon leads are provided for the transistors indicates the importance of this fact. The emitter leads in particular should be as short and direct as possible. An emitter resistor should not be used with balanced-emitter transistors since this is already provided internally. For some other types of transistors where insufficient protection is provided against load mismatch a small emitter resistor may be used. However, this resistor will reduce both power gain and power output.

Bypassing is critical in a high-power transistor transmitter due to the very low impedances involved. The best approach to bypassing power leads is multiple capacitors. A good technique is to use a feedthrough capacitor with other capacitors in parallel

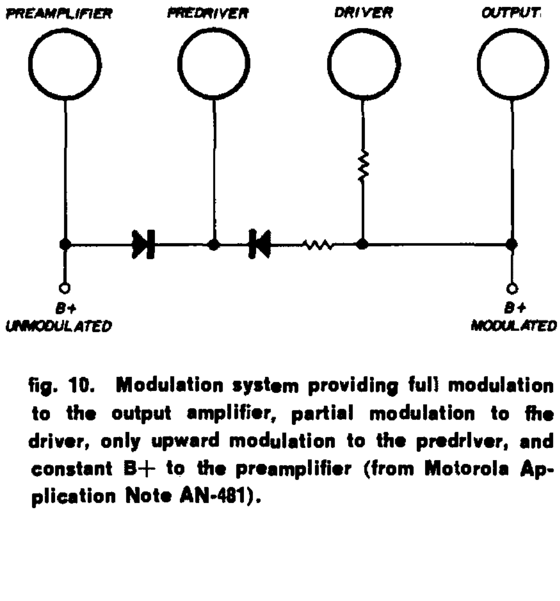



**fig. 10. Modulation system providing full modulation to the output amplifier, partial modulation to the driver, only upward modulation to the predriver, and constant B+ to the preamplifier (from Motorola Application Note AN-481).**

with it. For example, a 1000-pF feedthrough with a 0.1 μF disc ceramic capacitor and a 10-μF electrolytic capacitor in parallel helps assure good bypassing. (But don't use too much capacitance if you're applying audio for modulation.)

A good material for the chassis of a transmitter is copper or brass plate, or copper-clad printed-circuit board. If printed-circuit board is used be sure that an adequate heat sink is provided for the transistors. With these materials, components can be soldered directly to the chassis, assuring good grounds.

The input of each transistor should be isolated from its output as much as possible; in some cases, a shield may even be necessary where high gains are used.

The chokes used in a transistor transmitter should not have high Q; low-Q chokes help avoid many problems. If a high-Q choke is used in the base lead, for example, the transistor can take off at lower frequencies. Ferrite-core chokes are excellent in many cases. Ferroxcube VK-200 chokes are often recommended. Another approach is to use a couple of ferrite beads in series with another choke or even in series with just a small resistor or a piece of wire. In most cases, some experimentation is necessary to determine the best kind of choke. It's often a good idea to put a small resistor (10 ohms or so) in parallel with the base choke.

The coils and capacitors that are used in the collector circuit should be suitable for the high circulating currents. Don't forget that in a transistor transmitter currents are often many amperes and even a very small dc resistance can cause high losses.

One other problem with any type of vhf equipment, and one that is not well recognized by many amateurs, is the fact that resistors and capacitors have different values at high frequencies than they do at the frequencies where they are measured. For example, a 100-pF silver-mica capacitor can have a much higher capacitance at 2 meters. Unfortunately, most hams do not have facilities for measuring capacitance accurately at high frequencies.

If you have access to a good vhf bridge or a slotted line you can determine the actual value of a capacitor at the frequency of interest. Lacking this you may be able to use air variables; their capacitance varies much less than silver mica and ceramic capacitors.

In most cases it's possible to avoid resistors in places in the circuit where they are subjected to rf. This can be accomplished by careful circuit design.

One other important consideration in transmitter construction is the use of a low-pass filter in the antenna lead, or even better, a bandpass filter. This is necessary in vacuum-tube transmitters to avoid interference with tv sets and other communica-

tions. It's even more important in a transistor transmitter where the circuits tend to have lower Q.

## adjustments

A few hints for testing a transistor transmitter: rule number one is not to apply any power to a stage unless it's properly loaded. This means a dummy load suitable for the power level you are using. Light bulbs are not satisfactory; a Heathkit Cantenna, lossy coax cable or other good 50-ohm load is.

It's also a good idea to reduce power when you first tune up a transmitter; half voltage is enough. Adjust the tuned stages to approximate resonance if it's practical, since applying drive to a transistor without tuning its output circuit can cause problems. Probably no damage will result, though, if collector voltage is not applied to the transistor. The very low impedance of the base circuit makes it very difficult to develop enough voltage across it to blow out anything.

The usual way to tune a cw transmitter is to adjust it for maximum output with a wattmeter or dummy load and field strength meter. A better way is to look at the output on an oscilloscope. This can be done either with a direct connection to the plates of the oscilloscope, or with a mixer that will transform the high output frequency down to a frequency where your scope is usable. The mixer for this application does not need to be very complex. It's sometimes possible to use a receiver in this way if you're sure you're not overloading it.

It's a good idea to listen to the transmitter on your receiver at the output frequency. This will let you hear if any weird oscillation shows up. However, to have realistic results make sure that your receiver is not overloaded. A typical multiconversion vhf receiving system is very susceptible to overloading and all sorts of images. A simple diode detector and amplifier is probably more satisfactory for this application than your high-gain, low-noise converter.

Adjusting an amplitude-modulated transmitter is more difficult. Here you should tune for maximum upward modulation and least distortion, rather than simply maximum power output. The two rarely correspond. Here again, looking at the signal on a scope and listening to it are imperative.

Linear amplification is the most difficult of all. Here you should tune for minimum distortion. A scope is necessary; a spectrum analyzer is very useful if you can get one. If you're not careful with a linear amplifier, particularly in single sideband service, you may end up with a very high distortion and many spurious outputs.

In adjusting a transistor transmitter it's a good idea to use a regulated power supply, at least for initial adjustments. Most transistors are very sensitive to changes in supply voltage and you will get inconsistent results if your power supply voltage varies much.

## conclusions

This article has described the present state of rf power transistors and how they can be used in ham equipment. It has not gone into great depth in any subject; however, the list of references provide more information on the design and use of rf power transistors. Although rf power transistors are still relatively expensive, they are practical and should be carefully considered for use in your transmitting equipment.

### references

1. Frank Davis, "Matching Network Designs With Computer Solutions," Motorola Application Note AN-267, Motorola Semiconductor Products, Inc., Box 20924, Phoenix, Arizona 85036.

2. Roy Hejhall, "Systemizing RF Power Amplifier Design," Motorola Application Note AN-282, Motorola Semiconductor Products Inc., Box 20924, Phoenix, Arizona 85036.

3. RCA Silicon Power Circuits Manual, Radio Corporation of America, Electronic Components and Devices, Harrison, New Jersey.

4. John G. Tatum, "Vhf/Uhf Power Transistor Amplifier Design," Application Note AN-1-1, ITT Semiconductors, 3301 Electronics Way, West Palm Beach, Florida 33407.

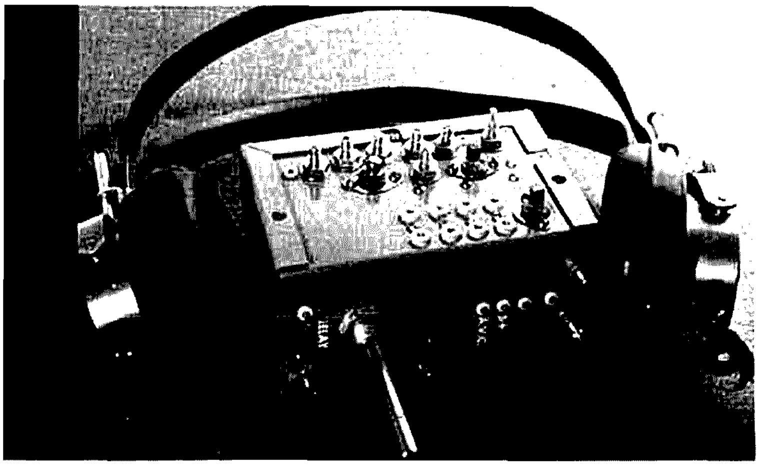
5. Frank Davis, "A 30-W, 175-MHz Power Amplifier Using PNP Transistors," Motorola Application Note AN-477.

6. Dick Brubaker, "A Broadband 4-Watt, Aircraft Transmitter," Motorola Application Note AN-481.

7. Roy Hejhall, "A 25-W, 175-MHz Transmitter for 12.5-Volt Operation," Motorola Application Note AN-503.

8. Dick Brubaker, "A 13-W AM Aircraft Transmitter," Motorola Application Note AN-507.

**ham radio**

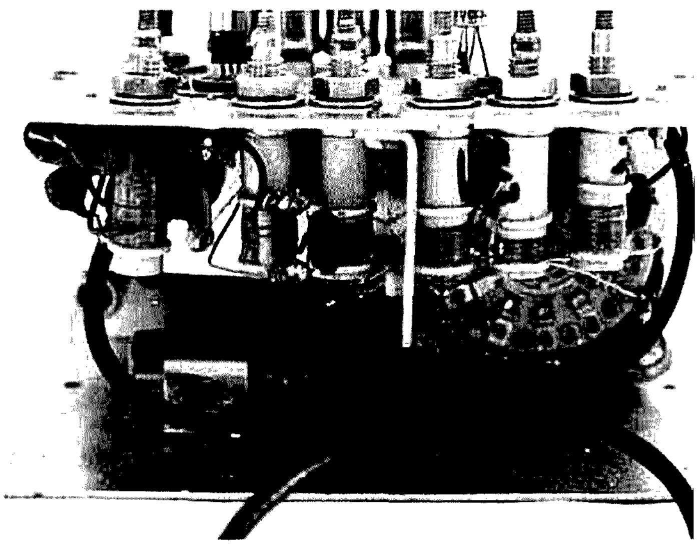


# second-generation fet converter for 10 to 40 meters

This bandswitching converter has improved stability and is much smaller than its prototype

Mike Goldstein, VE3GFN, 22 Kingswood Road, Toronto 13, Ontario, Canada

**This 40- through 10-meter** converter, was built as a design project for the 18th edition of the **Radio Handbook.**[2] Perhaps the most startling impression this converter offers it its small size. It would be difficult indeed to build a bandswitching converter covering this range into a smaller enclosure.

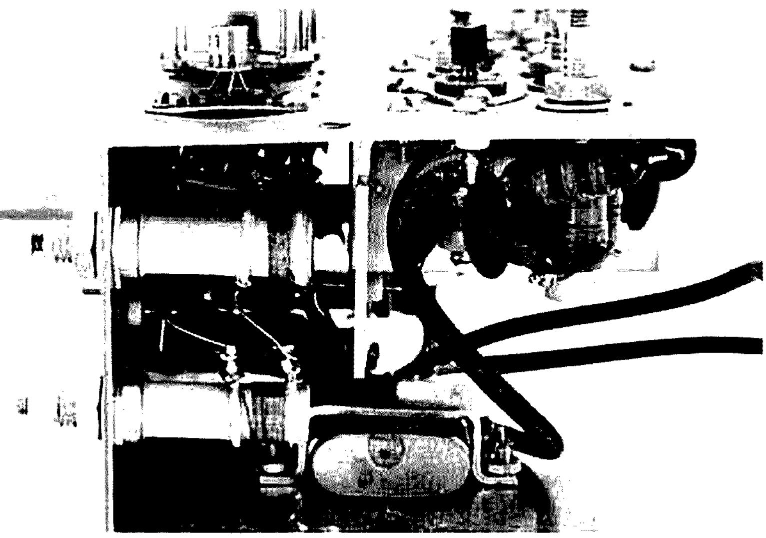


## design features

The design (fig. 1) was based on experience obtained from the previous model. This converter uses the simple bandswitching method of adding capacitance or inductance to a tuned circuit. An unneutralized rf amplifier is used, and no tuned circuits are switched in the oscillator.

The converter is less than one-quarter the size of its predecessor; however, it doesn't contain a power supply although it does contain its own change-over relay, which switches the antenna between converter input and input to the i-f strip. A dual-gate mosfet is used in the rf amplifier. This transistor has a feedback capacitance less than 0.03 pF, which means improved stability over the previous converter design.

The dual-gate transistor also allows this converter to be controlled by agc voltage, which is supplied to the mosfet second gate. Minus 1 volt is sufficient to cut off the transistor. The converter is completely broadband; the necessity for peaking tuned circuits on each band has been eliminated. So, the construction of a "second generation model has been justified—better stability, easier operation, agc control and much smaller size.



**Layout of oscillator output coils and mixer stage. Note change-over relay on bottom lip. Also note Teflon feedthrough terminals on shield section.**

The converter receives amateur bands through 10 meters and converts them to 80 meters merely by turning the bandswitch; no tuning is necessary. Input and output impedances are small, allowing the converter to be connected through random lengths of coaxial cable. Tuned circuits allow the complete 10-meter band to be covered, but only 500 kHz at a time, depending on the oscillator crystal frequency.

The rf amplifier is a common-source amplifier, unneutralized, designed basically as a 20-meter amplifier. To allow it to pass other bands, capacitance or inductance is switched across the tuned circuits. These tuned circuits have a 500 kHz bandpass on each band. On 10 meters, some loss in gain will be noticed if the 10-meter circuits are not retuned as crystals are changed to cover the different parts of the band. This method of bandswitching makes alignment very simple; only half the usual number of switch decks are necessary. Stability is assured by (1) using a transistor with a very small feedback capacitance, (2) using ferrite beads on amplifier input and output leads to reduce parasitic oscillation, and (3) eliminating amplifier tuning after initial adjustment.

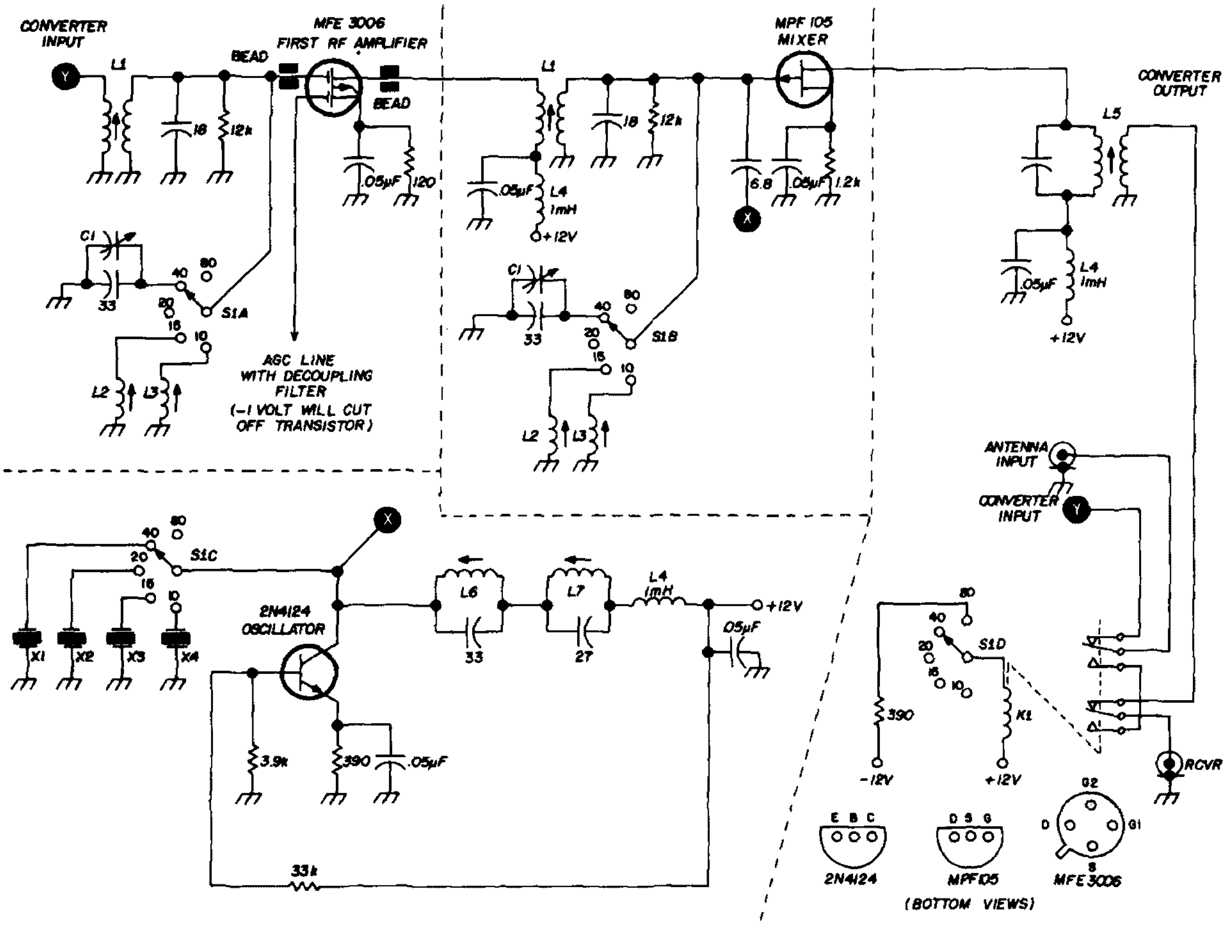
The mixer is also a common-source amplifier: a jfet with both injection inputs applied to its single gate. This is a reliable and simple circuit. It's identical to that of the mixer in the original version.

The oscillator uses a bipolar transistor that has a rather interesting output circuit. Output impedance, offered by rf choke L4, increases in reactance as frequency increases. Output voltage is thus increased to compensate for any high-frequency losses.

The crystal for the 10-meter band will probably be an overtone type. A 24.5 MHz tuned circuit is included in the output to pick off the proper overtone. This tuned circuit is a short circuit at all but its resonant frequency. A similar tuned circuit was also included for 15-meter operation after initial tests indicated the presence of a 20-meter image. This circuit affects operation only when the oscillator is switched to 15 meters.

## construction

Because of its size, construction of the converter may appear difficult. However, it's




**C1** 10 - 60 pF piston (Voltronics TM-60C)

**L1** 24 turns no. 32 closewound on Cambion 2022-2 form 4 μH, Q = 50. Link is 5 turns on 32 on cold end

**L2** 20 turns no. 32 closewound on Cambion 2022-3 form 3.4 μH, Q = 50

**L3** 11 turns no. 32 closewound on Cambion 2022-3 form 1.4 μH, Q = 50

**L4** Rf choke, 1 mH, 35 mA (Superex M-10)

**L5** 40 turns no. 32 closewound on Cambion 2022-2 form 11 μH, Q = 50. Link is 10 turns no. 32 on cold end

**L6** 10 turns no. 32 closewound an Cambion 2022-3 form. Resonates with 33 pF at 24.5 MHz.

**L7** 15 turns no. 32 closewound on Cambion 2022-2 form. Resonates with 27 pF at 17.5 MHz

**S1** 2-section, 6-pole rotary switch (Centralab 2021)

**X1** 3.5 MHz crystal

**X2** 10.5 MHz crystal

**X3** 17.5 MHz crystal

**X4** 24.5 MHz crystal

**fig. 1. Schematic of the second-generation fet converter. Dual-gate igfet rf amplifier allows improved stability and agc control.**

one thing to look at the photographs and see many components crowded into a small space; it's something else to have an empty box into which must be wired some very small components. With care and planning construction should proceed quite smoothly. While the converter is easy to build, mistakes will be extremely difficult to correct. Each step should be carefully checked before proceeding further.

The photographs show the layout. If time is taken to position components in each stage before wiring (including the shield partitions between stages) everything will fit quite neatly.

Looking at the photos, the three coils at the right rear are the rf amplifier coils; those at the left rear are the mixer coils. The

oscillator coils are at the front right, and the 40-meter piston capacitors are on either side of the bandswitch.

The photograph of the interior shows parts of the interstage shielding between the rf amplifier and the mixer. The change-over relay is mounted on the bottom lip of the converter. (This relay is from a military transponder and is about the size of a crystal; however, any dpdt relay that will fit will do.)

The rf amplifier should be wired first, after first drilling holes for *all* major components, including shields. The shields should be preshaped, and the push-in feed-through terminals should be installed before the shields are mounted. Each coil should be checked for continuity, and each tuned circuit should be grid-dipped as each bandwitch section is wired. This should be done *before* the transistors are plugged into their sockets.

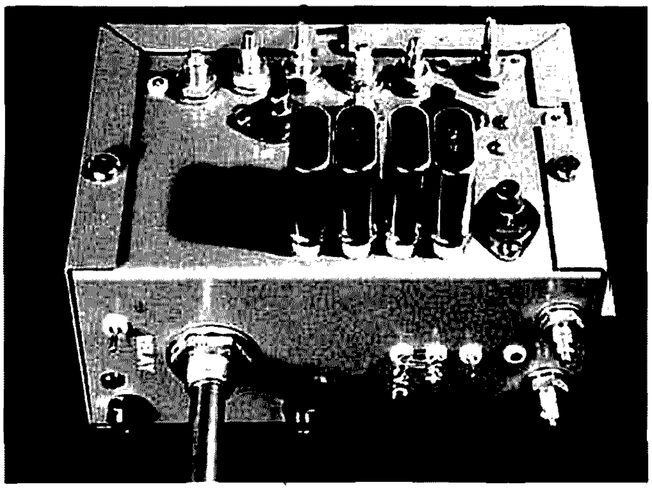
A very small soldering iron should be used, with a low-melting-point solder. Very narrow needlenose pliers are a must.

Miniature coaxial cable is used to route the signals around the relay. Care should be taken not to overheat this cable during installation. To limit power drain, the relay draws power only when the bandswitch is turned to the 80-meter position.

## alignment

Alignment is quite simple and can be done without any instruments except a receiver. Switch the converter to 20 meters

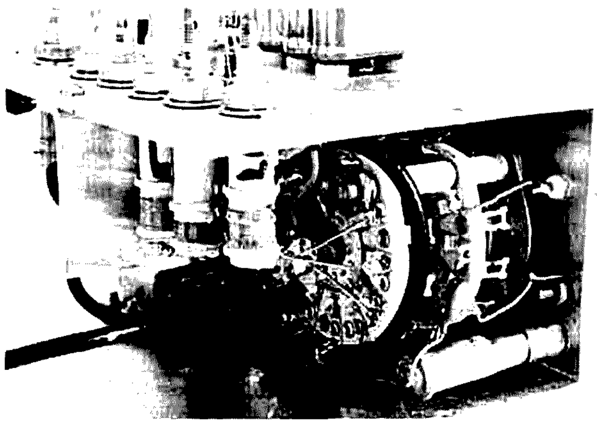
**Four-band fet converter Mark II.**



and the receiver to the middle of the 80-meter band. Connect an antenna to the converter, and peak the 20-meter tuned circuits on a received signal, starting with the mixer. Switch the converter to 40 meters, and peak the piston trimmer capacitors for maximum received signal, mixer first.

Now to 10 meters. The alignment here is a bit more involved, as the mixer tuning and the oscillator tuned circuit will both

**Rf-amplifier section. 40-meter piston trimmer is shown in front, next to bandswitch.**



affect the output. Find the correct beat frequency first with the mixer coil, adjust the oscillator coil for a peak in signal, then adjust the rf amplifier for maximum output signal. Now switch to 15 meters, and repeat the 10-meter sequence with the 15-meter circuits. If any sign of a 20-meter image appears on 15 (an ssb signal will appear to be on the "wrong" sideband), adjust the 15-meter oscillator tuned circuit to reject the image.

This completes the alignment of the converter. As for your severe case of eye strain, you might consult the "ham notebook" item on page 67 of the August, 1969 issue of *ham radio*.

### references


1. M. J. Goldstein, VE3GFN, "Bandswitching Fet Converter," *ham radio*, July, 1968, p. 6.
2. W. I. Orr, W6SAI, "The Radio Handbook," 18th edition, 1969, Editors and Engineers, Ltd., New Augusta, Indiana 46263.


**ham radio**

# random-length antenna couplers

## Solutions to the antenna-matching problem using proven switching circuits

John J. Schultz, W2EEY, 40 Rossie Street, Mystic, Connecticut 06355

**With a few basic components** you can build an antenna coupler that will couple a transmitter to almost any antenna length on any band. The key to this coupler is to use all the possible interconnections between the coupler components.

When using a random-length wire for an antenna, either in a portable or fixed station, the usual procedure is to build a pi-network coupler to match the antenna to a transmission line or transmitter designed to work into a fixed antenna impedance (usually 50-70 ohms).

The pi-network is capable of matching a wide range of impedances, including some reactive antenna loads, to a fixed-impedance transmission line or transmitter—once the proper component values for the network are determined. However, it's not always the most economical or efficient circuit, especially if a random length of wire is used on several amateur bands. Also, if the antenna wire length changes drastically, as in portable operation with unpredictable geographical restraints, the fixed pi-network can require considerable readjustment before proper antenna tuning results.

The problem is that one often tries to put too much flexibility into a circuit using components of a fixed (finite) value. Changing component values beyond their normal variable range is impractical. Therefore, the only other solution is to alter the circuit for greater matching flexibility.

This article explores the range of coupling circuits using up to three components (two capacitors and one coil or two coils and one capacitor). The variety of possible circuits will surprise those used to the familiar pi-network as being the only useful three-component matching network.

These circuits are presented as impedance-matching devices only. They vary greatly in their ability to attenuate harmonics (the pi-network ranks high in this category). However, assuming you have a reasonably harmonic-clean transmitter and are concerned only with coupling for maximum power transfer to a random-length antenna, many of the circuits should be very useful.

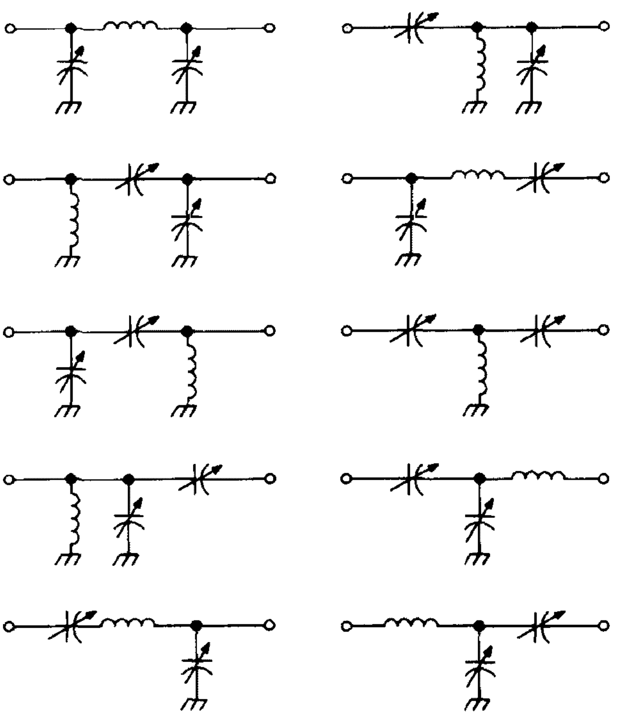


**fig. 1. Matching networks that use one inductor and two capacitors.**

## circuits containing a single inductor

**Fig. 1** shows circuits using a single inductor and two capacitors. Notice that some are simply others "turned around." They should not be disregarded, however, because either an impedance step-up or step-down between antenna and transmitter or vice-versa may be necessary. Each circuit represents a specific matching network possibility.

The range of reactive impedances a circuit can match depends on frequency and the range of the variable components. Often it will be found that more than one circuit will easily match the same antenna load to a transmitter. Deciding on the best circuit is covered later.

## multicircuit switching

An antenna coupler can be built with sufficient switching capabilities to use all ten of the possible combinations shown in **fig. 1.** All circuits except one can be constructed with one side of one of the variable capacitors connected to ground. This greatly simplifies construction since only one "floating" variable capacitor is required.

**Fig. 2** shows a circuit that allows switching for 9 of the 10 circuits shown in **fig. 2.** Practical component values for a typical 80-10 meter coupler are also noted on the diagram. The switch has only 5 positions; the other coupler circuits are formed by reversing the coupler connections. Thus, for each switch position, the coupler has to be tried both forward and reversed. An internal reversing switch would probably be worthwhile for portable use where the antenna length or band is frequently changed.

## component ratings

The voltage rating of the capacitors, switch insulation, and coil depend a great deal on the use to which the coupler will be put. Transmitter power level and the range of antenna reactances are significant. The components in **fig. 2** should work well with a 100-150 watt transmitter with almost any antenna length. For higher power, capacitors with larger spacing and a heavy coil will be necessary if very short antennas are to be matched without arcing. The cost of such a coupler, unless surplus components are used, can be very high for transmitters having more than a few hundred watts output. It might be better in this case to use a loaded antenna.

## construction

The construction of the coupler entails no particular precautions. Since harmonic attenuation is not a consideration, the coupler can be installed in a plastic container. This immediately solves the problem of ground-floating variable capacitors. The switch should be wired with the same size wire as the coil. Leads on the switch sections should be as short as possible. The wiring should present no problem if the scheme shown in **fig. 2** is followed.

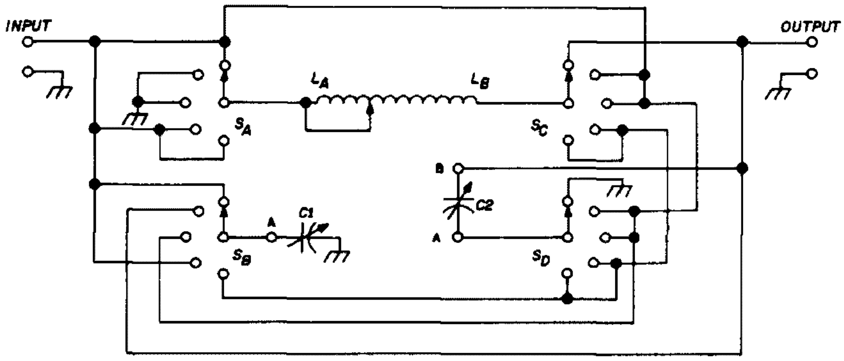



**fig. 2. Nine of the ten circuits shown in fig. 1 can be formed with this circuit (the tee network with the two floating capacitors is the exception). Capacitors C1 and C2 are 250-pF air variables (Hammarlund MC-250-M); the inductor is a B&W 3900, 5 to 6 inches long, tapped with a spring clip.**

## dual-inductor circuits

Matching circuits using two inductors and one variable or fixed capacitor are shown in **fig. 3.** These are the equivalent of the single-inductor, two-capacitor circuits of **fig. 1,** but they do have some different features.

With extremely short antennas (one-eighth wavelength or less), some of the **fig. 3** circuits will give superior coupling efficiency—provided the inductors have low ohmic losses. This is very important, as discussed below.

The high circulating currents in a network matching an extremely short antenna could be 20-50 amperes. This would produce serious $I^2R$ losses. With a fixed antenna length, the only way to reduce these losses is to construct the coupler with components having the lowest possible losses for the frequencies involved.

Antenna couplers having an extremely wide range of impedance-matching capabilities can be made using the circuits of **fig. 3** if each inductor has a maximum value of 10-20 $\mu$H and the variable capacitor has a range of 350 pF or more. This should permit matching a 16- to 20-foot or rod wire on 80 meters without difficulty. The cost of such a coupler can be fairly high if new commercial components are used. If you use surplus components, such as the capacitors and roller inductors in BC-191 and BC-375 units, very good couplers can be constructed at moderate cost. Alternatively, the inductors can be made from 1/8- or 1/4-inch copper rod or tubing to provide a 10-20 $\mu$H inductance.

In any case, when dealing with extremely short antennas, I'd suggest a hard-wired interconnection of the coupler components rather than a switching arrangement. It's difficult, but all-important, to remember that with such an antenna you're dealing with very high currents with even a medium-power transmitter. With circulating currents of 40 amperes, for instance, it takes only **0.1 ohm** in the inductors or connections to the matching network to throw away 160 watts of power in heat loss.

The contact resistance of many simple switches exceeds this 0.1-ohm value, and this fact alone explains why many well-constructed couplers seem ineffective when used with extremely short antennas. (Mobile enthusiasts take note.) The advantage of low ohmic resistance components in a coupler

for extremely short antennas cannot be overemphasized. That's why military couplers use oversized components for moderate power levels.

## tuning and adjustment

If you have only a pi-network available as an antenna coupler, the usual procedure would be to adjust it for minimum standing-wave ratio (swr). Indeed, this would be the

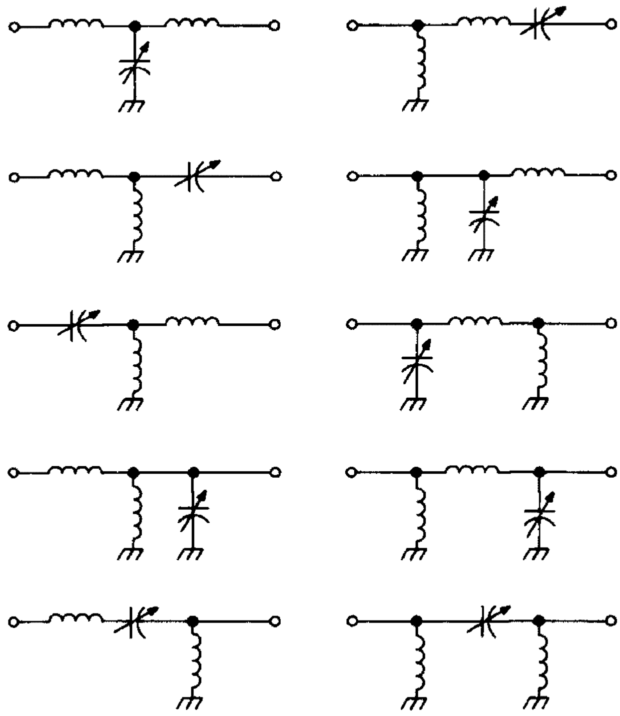


**fig. 3. Matching networks that use one capacitor and two inductors.**

only possible matching adjustment when tuning up a transmitter with a pi-network antenna coupler.

When using a coupler of the type in **fig. 2** with random-length antenna wires (30 feet or more), note that several coupler forms on the same band might produce a minimum swr. In each case all coupler component values should be adjusted to achieve a minimum swr when the coupler is set for a specific circuit form. The same situation will often occur when using the coupler forms shown in **fig. 3.**

## swr and field strength

Minimum swr doesn't necessarily mean maximum radiated power. Minimum swr can result from transmitter power completely transferred to the antenna, completely absorbed in the coupler network, or any condition in between. The only way to determine what is taking place is with a field-strength meter.

The field-strength meter should be placed away from the coupler's immediate field. It doesn't matter which coupler circuit is tried first, as long as the sensitivity control on the field-strength meter isn't touched once a reading has been established for a particular coupler circuit. Other coupler circuits should be compared against the first, which is established as a reference, until one is found that produces the greatest field intensity.

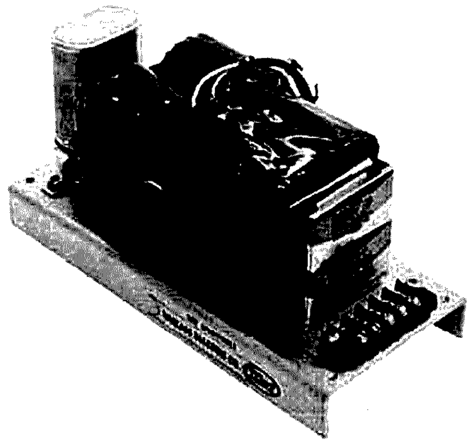
After you determine which coupler circuit produces the lowest swr with greatest field strength on a given band, you can log the coupler settings for changing bands quickly. The time spent in initial setup to determine the optimum coupler circuit with each different antenna or ground connection will be worthwhile.

The matching networks shown can be used on any band, but the unconventional types will find their greatest application on the high-frequency bands when used with odd-length antennas. For simple situations where an impedance match is desired and no reactive components are involved, a simple L or pi-network generally suffices. The latter circuit also has some harmonic suppression as well.

## grounds

A good ground connection is always desirable when using a random-length antenna, particularly if the antenna is less than a quarter wavelength long. In portable work, care should be taken to check the radiated field strength both with and without the ground connection. It's possible to have a ground connection that will actually reduce the radiated signal. You could lay a wire along the ground as a substitute for a radial system. In any event, the installation should be checked with a field-strength meter, as described above.

**ham radio**



# power supply protection
## for your solid-state circuits

Carl C. Drumeller, W5JJ, 5824 N. W. 58 Street, Warr Acres, Oklahoma 73122

**Many amateurs have learned,** by unpleasant experiences, that transient voltage spikes and solid-state devices make a disastrous combination. A number of schemes have been advanced for limiting (if not eliminating) the deadly effect of even a moment's overvoltage. These vary from selenium clipping diodes to RC spike absorbers. Most work; some don't, which accounts for the thriving business in replacement diodes. Usually, the failure can be traced to omission of a precaution against an unexpected source of overvoltage.

A failure-proof device is now on the market, designed to give peace of mind to the user of solid-state devices. It's sold under the trade name Paraformer™ (derived from parametric transformer). A number of desirable functions are combined in this unique transformer. It will regulate changes in line voltage and changes in load. You can feed it sine waves, square waves, or waves with superimposed modulation; in each instance, the output is a sine wave (or nearly a sine wave). You can hit it with a voltage spike running into thousands of volts without affecting its sine wave output. It also attenuates (by 50 dB) all types of noise up to one megahertz. With the addition of electrostatic shielding (normally not required), its noise-reduction capability is effective far into the megahertz region.

### voltage surges

A significant feature is that turn-on and turn-off time is not instantaneous. Approximately 6 cycles are required for the voltage to build up from zero to 117 V and for the voltage to decay to zero when the input is switched off. Think what this means. No longer do you have to worry about elaborate circuits to sense when a waveform is passing through zero and break the circuit at that point. You can turn on a simple switch at any time with full confidence that the voltage will increase slowly enough to avoid any trace of a spike. Quite probably you can delete surge-limiting resistors between rectifiers and filter capacitors. At least the initial surge shouldn't be too drastic. What happens after that depends on how much of the charge you drain out of the capacitor between charging pulses.

### how it works

Conventional transformers used at power frequencies consist of two coils (primary and secondary) wound on a common core of

magnetic material. A current flowing in the primary creates a magnetic field consisting of lines of force (magnetic flux) that link with the turns of the secondary. If a load is connected to the secondary, the mutual inductance of the two coils will cause a voltage to be induced in the secondary.

The common transformer is a fairly efficient energy-transferring device, but it has one serious disadvantage. Because it operates on the principle of mutual inductance, it cannot discriminate between noise and transients appearing on the input voltage. It transforms noise and spikes just as efficiently as the applied voltage.

The Paraformer also has a primary and secondary coil wound on a magnetic core.

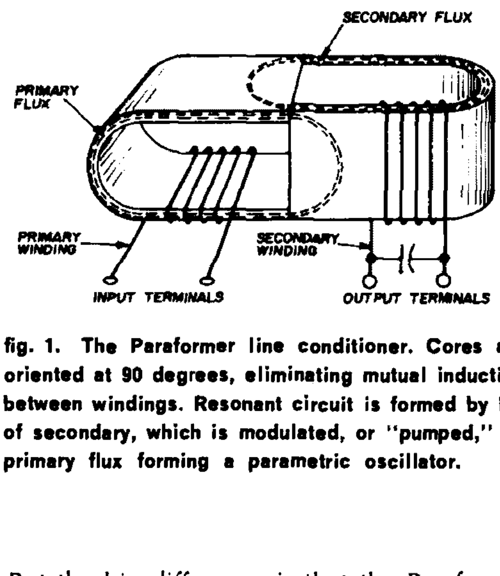



**fig. 1. The Paraformer line conditioner. Cores are oriented at 90 degrees, eliminating mutual induction between windings. Resonant circuit is formed by LC of secondary, which is modulated, or "pumped," by primary flux forming a parametric oscillator.**

But the big difference is that the Paraformer is designed to have zero mutual inductance between primary and secondary. This is achieved in a novel manner, as shown in **fig 1.**

The core consists of two elements positioned at 90 degrees with respect to each other. Note, however, that both coils share some of the common core material.

The opposition to flux linkage in a magnetic circuit is called reluctance. In the Paraformer, the reluctance of the secondary winding, and hence the energy transfer, is varied in a controlled manner. This is known as "pumping" the secondary. In **fig. 1,** note that the secondary coil and capacitor form a parallel resonant circuit at some frequency, f. If the reluctance of the secondary, and hence the flux coupling, is varied at 2f; and if the pump frequency is **exactly in phase** with that of the input voltage; then a phase-locked oscillator results.

The energy supplied at 2f, which pumps the secondary, controls circuit impedance in a manner to increase the power at f. The important difference between the Paraformer and other transformers, including the ferroresonant transformer, is that energy is transferred between primary and secondary at frequency, f, without transferring voltage spikes or noise that may appear on the input voltage. This is because energy transfer is obtained by a controlled input (pump) signal that's **phase locked** to the output. Thus output power is noise free and remains constant, even when input power momentarily drops out. Further details on the Paraformer are available in reference 1.

These line conditioners, as they're called, don't cost much more than an ordinary regulated power supply. But what a difference! For example, the PEC-60 (shown in the photo) has features unheard of in an ordinary regulated supply. Protection against over-voltage without external sensing devices is probably the most important feature for amateur work. If you carelessly plug it into a 220-volt line, its output instantaneously drops to zero. No blown fuses; no ruined transistor circuits. If power is continuously demanded by the load, output ceases, because the parametric oscillator won't function. Conversely, if input voltage is low, the parametric oscillator will not receive enough power to overcome circuit losses. Consequently, no oscillation will occur, and output voltage drops to zero. Between extremes of under- and overvoltage, the Paraformer continues to regulate at its design center.

## references

1. Wanlass Electric Co., "Recent Technical Articles on Parametric Power from: EDN (Electrical Design News), Engineer/Scientist, Electromechanical Design," Wanlass Electric Company, 2175 South Grand Avenue, Santa Ana, California 92707

**ham radio**

# logarithmic speech processor

An addition to your phone rig that will increase average power by 8 dB without distortion

Lee M. Richey, WA3FIY, RD 1, Franklin, Pennsylvania 16323

**Many articles have been published** describing speech clippers, rf clippers, automatic level-control circuits, and other methods of obtaining higher average transmitted power without creating spurious radiation. The speech processor described here also does this, but a different method is used.

The human speech waveform has a very low average-to-peak power ratio. The peaks are several times greater than the waveform's average amplitude, and the peaks are of very short duration. The power in these peaks is a very small portion of the waveform's total power because of the short durations of the peaks. If we can in some manner control the peaks to increase the average-to-peak ratio in the waveform, we will add little distortion and much power to the transmitted signal.

At this point, I suppose I could become involved in the argument between high-fidelity and communications-quality enthusiasts, but I won't go into that now. Let me say that the speech processor described here can be adjusted to suit almost every taste in signal quality and will, in every case, increase the average transmitted power.

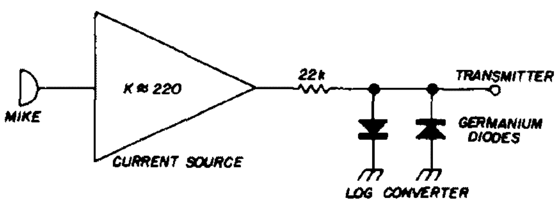
## speech clippers: pro and con

Most speech and rf clippers work on the same basic principles. When the speech or rf waveform exceeds a predetermined amplitude, the remaining waveform is clipped from the output. This system has been used

in many commercial and home-built transmitters, but it has two disadvantages. First, the clipping level is usually fixed. This means clipping varies with how loud you speak into the microphone. This can have a drastic effect on the amount of distortion in the output. When speaking loudly into the microphone, the average amplitude may exceed the clipping level. Thus, most of the speech intelligence is eliminated from the waveform. This makes the signal loud but very difficult to copy.[1]

The second disadvantage is that harmonic distortion is created by clipping. Since clipping results in an abrupt change in the rate of rise of the speech waveform, many odd-order harmonics are generated. The clipped portion of the waveform resembles a square wave, and a perfect square wave contains an infinite number of odd harmonics of the fundamental frequency.

**fig. 1. Basic circuit of the logarithmic speech processor. A high-gain amplifier provides a current source to drive a pair of diodes connected back-to-back.**




These harmonics can cause an excessively wide signal. They can be eliminated by a low-pass filter at the clipper output, but this adds to the size and cost of the unit. Filter-type single sideband transmitters limit the transmitted signal bandwidth automatically, but the problem of the fixed clipping level remains.

Both of these disadvantages can be eliminated by the logarithmic speech processor, which eliminates the abrupt change in the waveform and eliminates the problems caused by the discrete clipping level.

## the logarithmic converter

With this system, the waveform's average-to-peak ratio is increased by the logarithmic conversion of the speech waveform. Its output voltage is proportional to the logarithm of the input voltage. Thus, the abrupt change in the waveform is eliminated, because the log conversion is continuous and smooth. No specific clipping level predominates to cause harmonic distortion. The basic circuit is simple and can be used with am, fm, or ssb transmitters without filters or other accessories (**fig. 1**). Logarithmic conversion is accomplished with a current source and a pair of germanium diodes. The characteristic curve for a common small germanium junction diode, the 1N34A, shows a smooth logarithmic response on voltage versus current (**fig. 2**).

## operation

The voltage across the diode is proportional to the log of the current through it, so a current source is necessary. This is a high-gain voltage amplifier, which drives the diodes through a high resistance. The amplifier output is about six volts p-p. With a 22 kilohm series resistor, this represents about 360 μA p-p available to the diodes. One diode conducts during the negative half cycle, and the other conducts during the positive half cycle, so each diode shares half of the p-p current.

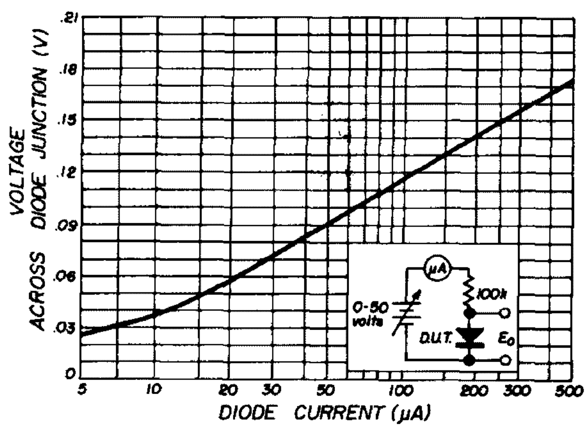


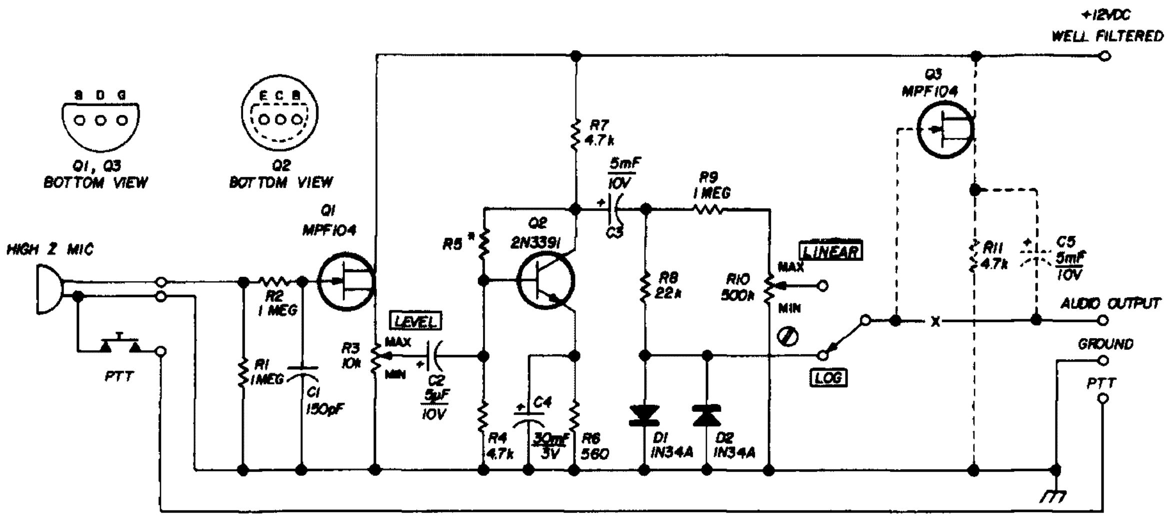

**fig. 2. Curve for a common germanium diode (1N34A). Logarithmic transition of output voltage versus input current is smooth and continuous.**

**Fig. 2** shows that the voltage across each

diode at 180 $\mu$A will be about 137 mV for a total output of 274 mV p-p. This is more than enough audio to drive most transmitters.

## the circuit

The high-gain amplifier consists of an fet source follower input, for high impedance,




*ADJUST R5 FOR COLLECTOR VOLTAGE OF Q2 TO BE SLIGHTLY OVER 6V. (START WITH ABOUT 18k)

**fig. 3. The logarithmic speech processor schematic. An fet source-follower output circuit (dotted lines) is recommended for use with low-impedance input equipment.**

and a common emitter amplifier, **fig. 3.** Output may be taken directly from the 1N34A's if the processor is to be used with equipment having high impedance input (250 kilohm or greater). If the circuit is to be used with low-impedance equipment, such as the SB-34 transceiver, a source follower output circuit (dotted lines, **fig. 3**) is recommended.

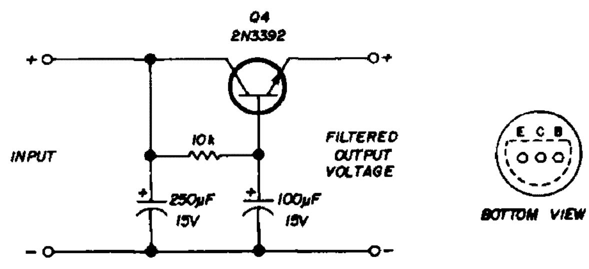
Power can be borrowed from other equipment, or it can be supplied by a battery. Power requirements are 9 to 12 Vdc at 10 mA. Very smooth dc is required. If necessary, the filter shown in **fig. 4** may be used for additional filtering.

## construction

The unit shown in the photographs was constructed on "Micro Vectorboard," which is a G-10 epoxy glass board with 0.042-inch holes spaced on 0.1-inch centers. Component leads are inserted through the holes, bent over, and soldered on the opposite side, printed circuit fashion (**fig. 5**). As with all high-gain, low-level circuits, leads should be kept short and direct; and input and output leads should be shielded as well as those to the log-linear switch.

The transistors shown on the schematic were used because of their low noise characteristics. Others may be used, of course, but keep in mind that the voltage gain of the circuit is over 200, and noise could be a problem with poor transistors. The unit shown in the photo was mounted behind a control panel with some other equipment, but there is no reason why it couldn't be mounted in its own small enclosure or within the transmitter or transceiver.

**fig. 4. Filter circuit for source voltage to ensure low ripple content.**



## adjustment

After completing the wiring, but before mounting the unit, apply power and adjust the value of R1 until the collector voltage of Q2 equals slightly greater than one-half the supply voltage. After mounting the unit in its enclosure, set the switch to the **log** position and the two potentiometers to their **min** positions.

Probably the best way to set the level control is to connect the output of the processor to a high-fidelity amplifier and listen through a pair of earphones. While speaking into the microphone, adjust the level control until the distortion just becomes objectionable. Turn the control back slightly, and you should be all set to go. Turn the switch to **linear** and, while talking into the microphone with the same loudness as before, adjust the linear control to approximately the same loudness or slightly less.

Now connect the processor to your transmitter microphone jack, and adjust the microphone gain control for proper transmitter operation. If you have a general purpose oscilloscope available, check the waveform at the collector of Q2 to be certain that this stage isn't flat topping. If it is, lower the level control slightly. If a high-fidelity amplifier and an oscilloscope aren't available, you'll have to depend upon on-the-air reports to find the proper level-control setting. The linear control should be set after the level control has been set, and the linear control should be adjusted for proper transmitter opeartion.

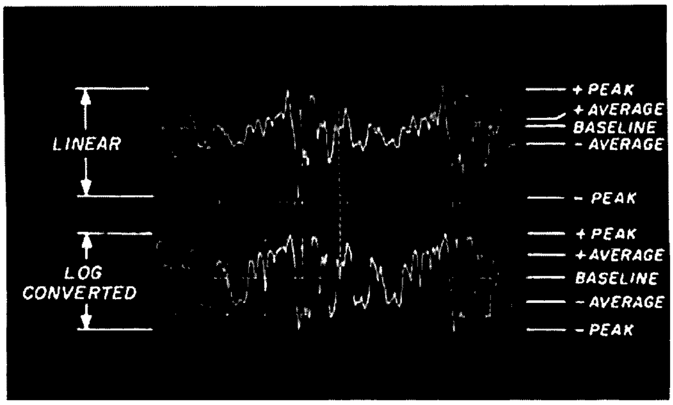



**fig. 3. Oscilloscope traces showing part of the word "hello". Upper trace shows waveform normally fed to transmitter; lower trace, after log conversion, shows an 8 dB increase in average-to-peak ratio.**

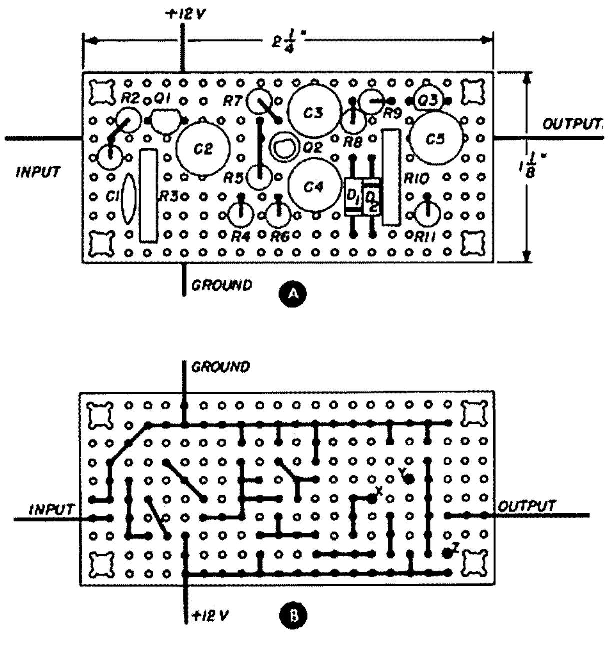



**fig. 5. Perforated board layout. Component side is shown in A; wiring side in B. Points x, y and z go to the linear-log switch.**

## operation

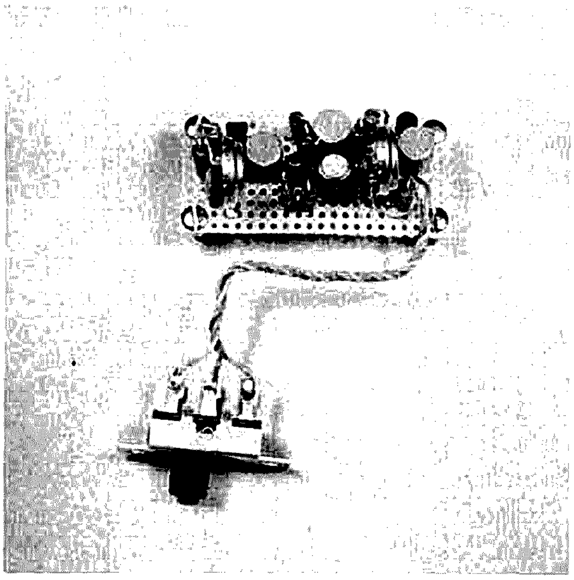
The plate meter of our transmitter should indicate a higher average current with the processor switch in the **log** position than in the linear position. Yet, the peak output as observed on a monitor scope should be no greater. (Caution should be observed, when operating near the legal power limit, not to exceed 1 kw dc plate power input.) The linear position bypasses the log conversion, and operation is normal in this position.

This unit works equally well with transmitters with and without alc. In sideband rigs with alc, it allows higher average plate current before the alc circuit takes control. In rigs without alc the processor allows higher average plate current before flat topping. In am and fm equipment it allows a higher average percentage of modulation, without overmodulation, and it does all this without any noticeable distortion.

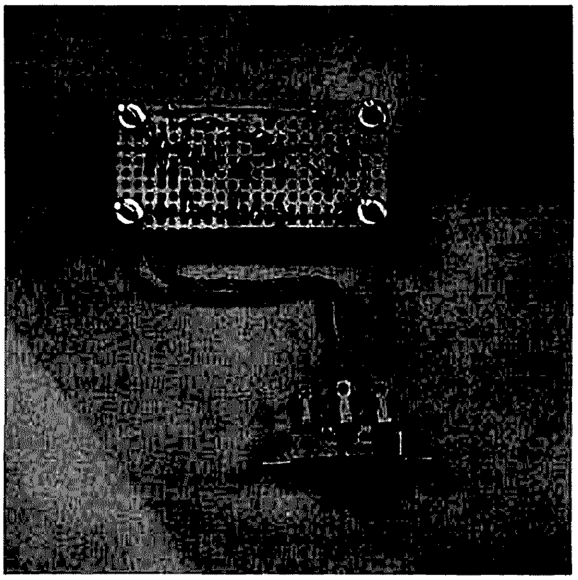
competing with much higher-powered equipment and the audio reports have all been favorable.

## test results

The oscilloscope traces of fig. 6 offer dramatic proof of what can be achieved with controlled speech processing. Both traces show the waveform of a portion of the word "hello." The upper trace shows that the



**Top view of finished speech processor ready for mounting. The switch in the foreground is the "linear-logarithmic" switch. Input to the circuit is at the left side; output at the right.**



**Bottom view of the speech processor. The wire leads of the components are inserted through the circuit board, bent over the leads of other components, and soldered together to produce the unit.**

average-to-peak ratio is about 1:5 for both positive- and negative-going voltages. The lower trace shows the same waveform after logarithmic conversion. Notice that the average-to-peak ratio is now about 1:2.

The improvement in average-to-peak voltage ratio, after log conversion, is 2:5. This represents an 8 dB increase in average power, and this does not include distortion products!

## some final thoughts

As with all such devices in this class, it must be remembered that since the average-to-peak ratio has been increased, the signal-to-noise ratio is lowered. Hum and noise in the microphone wiring, room noise, fans, blowers, and other extraneous noise will modulate the transmitter more than before. Therefore, care should be taken with regard to microphone wiring, placement and use.

This speech processor has been in operation for over six months now and has given good service. It has been used with an SB-34 transceiver barefoot with 65 to 70 watts dc input. I have been able to hold my own in

### reference


1. E. H. Conklin, K6KA, "To Clip or Not To Clip?" *ham radio*, April, 1969, p. 24.


**ham radio**

# proportional temperature control for crystal ovens

Design data and a practical circuit to ensure precise thermal control for crystal oscillators

J. A. Koehler, VE5FP, 2 Sullivan Street, Saskatoon, Saskatchewan, Canada

**When building high-stability** crystal-controlled oscillators, it's necessary to ensure that the crystal operates at a constant temperature. One of the best ways of doing this is to put the whole oscillator inside a tin can and bury it somewhere. The ambient temperature five or six feet underground is remarkably constant. This method is excellent if you're prepared to put up with the inaccessibility, near impossibility of repair, and the problems associated with drilling holes in your walls.

Putting the crystal in a small electrically heated oven, regulated by a thermostat, has none of these disadavantages. Such ovens often can be found in junked or surplus equipment. However, thermostats are rarely adjustable, so the oven temperature is governed by whatever gods look upon surplus emporiums. The thermostat contacts carry all the heating current and therefore tend to get dirty and arc after much use. This causes noise, which may get into other equipment through the heating-current line.

Also, the temperature in these ovens varies cyclically as the heating element is alternately turned on and off. This causes a small periodic variation in oscillator frequency, which may be undesirable in critical applications.

## proportional temperature control

A proportional temperature control is one in which the heating current is varied smoothly under the control of a temperature sensor. The current is never switched entirely off or on, but is left on continuously; as the temperature rises toward the desired level, the current decreases. Finally, a situation is reached where the current is just sufficient to maintain the temperature at a constant value. If the outside temperature decreases, the controller increases the current until equilibrium is again reached.

Normally a thermistor is used as the temperature sensor and is connected to an amplifier that controls the heating element. There is, however, a simpler way that makes use of the leakage current in transistors and which is highly temperature dependent, so that a transistor may be used as the sensor.

## the circuit

**Fig. 1** shows a circuit that takes advantage of this principle. Q1 is a leaky germanium (Ge) pnp transistor and is the temperature sensor. Q2 is an amplifier, and Q3 is the heating element. Q3 is a power transistor able to handle the required heating current. To ensure that the heat generated by Q3 is distributed evenly throughout the oven, Q3 should be fixed to the oven walls if they are metal, or to a strip of metal extending throughout the oven if it is not metallic.

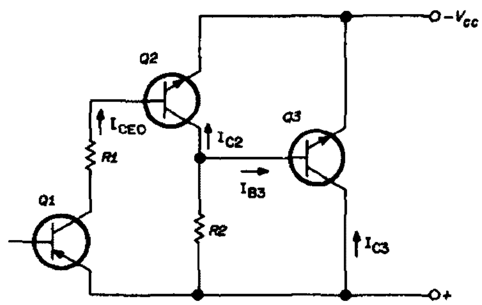



**fig. 1. Proportional temperature control circuit. Q1, a leaky germanium pnp transistor, is the temperature sensor; Q2 is an amplifier, and Q3 is the heating element.**

This circuit provides a simple, inexpensive, and precise way of controlling temperature. Perhaps the greatest advantage of the circuit is that it may be easily analyzed and may therefore be used to design a controller for virtually any desired temperature. In fact, for a given set of transistors and $V_{cc}$, the temperature is determined only by the value of R2.

## analysis

Q1's base is floating, so its collector current is $I_{ceo}$. $I_{ceo}$ is very dependent on temperature. For Ge transistors, it increases by a factor of ten for a temperature increase of 55° F.

Suppose you measure $I_{ceo}$ at room temperature (70° F) and call the measured value $I_1$. Then $I_{cbo}$ at any other temperature, T, will be

$$I_{ceo} = I_1 \, (10)^{\left(\frac{T-70}{55}\right)}$$

This current is fed to the base of Q2, so the collector current, $I_{c2}$, of Q2 will be determined by the common emitter current gain, $\beta_2$, of Q2 by

$$I_{c2} = \beta_2 \, I_{ceo} = \beta_2 \, I_1 \, (10)^{\left(\frac{T-70}{55}\right)}$$

Q3's collector current, $I_{c3}$, is determined by Q3's beta, $\beta_3$. Its base current, $I_{b3}$, is determined similarly, so that $I_{c3}$ is equal to $\beta_3 \, I_{b3}$.

Now, if Q2 were not connected to the base of Q3, $I_{b3}$ would be determined by resistor R2 and would be approximately equal to $V_{cc}/R2$. However, Q2 draws some current, so the current available to Q3's base is only

$$I_{b3} = \frac{V_{cc}}{R2} - I_{c2}$$

Substituting the value of $I_{c2}$ determined previously into this equation and multiplying it by $\beta_3$ yields the collector current to Q3:

$$I_{c3} = \beta_3 \left[ \frac{V_{cc}}{R2} - \beta_2 \, I_1 \, (10)^{\left(\frac{T-70}{55}\right)} \right]$$

Assume you've connected the circuit and apply $V_{cc}$ when the components are at room temperature. At that time, T = 70° F, so $I_{c3}$ will be

$$I_{c3}\,(\text{initial}) = \beta_3 \left[ \frac{V_{cc}}{R2} - \beta_2 \, I_1 \right]$$

The power dissipated by Q3 at that instant will be the product of $V_{cc}$ and this initial current. As the temperature rises, $I_{c3}$ will decrease; eventually the temperature will reach some final value, $T_f$, where the heat dissipated by Q3 is just enough to balance the heat losses. At this point, $I_{c3}$ will be

so small compared to its initial value that we can approximate it to zero. Then

$$\frac{V_{cc}}{R2} = \beta_2 I_1 \ (10)^{\left(\frac{Tf-70}{55}\right)}$$

This equation may now be used to calculate the value of R2 that must be used for the final temperature, Tf, of the oven. **Fig. 2** is a nomograph of equation 2. To design an oven controller for any temperature, simply calculate $V_{cc}/\beta_2 I_1$ for the transistors you intend to use, and use **fig. 2** to obtain the value of R2.

R1 is used to limit the dissipation of Q1. The leakage current in Q1 will cause some heating of the junction. If Q1 is to sense the temperature inside the oven, its dissipation should be as small as possible. A safe value of R1 is a value of kilohms equal to $V_{cc}$ in volts. For example, if $V_{cc}$ is six volts, R1 should be at least 6 kilohms. Q2 and Q3 should be silicon (Si) transistors. If they are Ge, the final temperature will end up somewhat lower than the design value, since leakage in Q2 will decrease in the base current available to Q3.

In an extreme case Q2 may even run away thermally and turn Q3 completely off (and perhaps burn out Q2). However, for temperature less than about 125° F, Ge transistors will perform satisfactorily. In any case, Q2 and Q3 must be the same type; either Ge or Si.

## measurement considerations

Uncertainties in the measured values of $\beta_2$ or $I_1$ cause surprisingly little effect in the final temperature. For instance, if your measured values of either $\beta_2$ or $I_1$ are in error by a factor of two, the final temperature will be within about 16° F of the design value. If you need a specific and precise temperature, make R2 variable and adjust it to get the desired temperature. Before building the circuit, use **equation 1** to calculate the initial collector current of Q3 to ensure that it does not exceed the rated maximum for the transistor. If it does, you'll have to use a heftier transistor.

I used this cicruit to control a small oven for the clock in a digital counter I'm building. The oven is about 1 × 1 × 2 inches. The design temperature was about 100° F. since the temperature here seldom exceeds that value. Q1, which is of uncertain origin, is a pnp Ge transistor in a TO-5 case. $I_1$ is 20 μA. $V_{cc}$ is 12 volts, so R1 is 12 kilohms. Q2 is a silicon planar transistor (probably one of the 2N3704 family). It has a measured beta of 140. Q3 is a TIP-24, with a measured beta of about 80. R2 is 1 kilohm.

Substituting these values into equation 2 results in a predicted final temperature of 102° F. The measured final temperature was

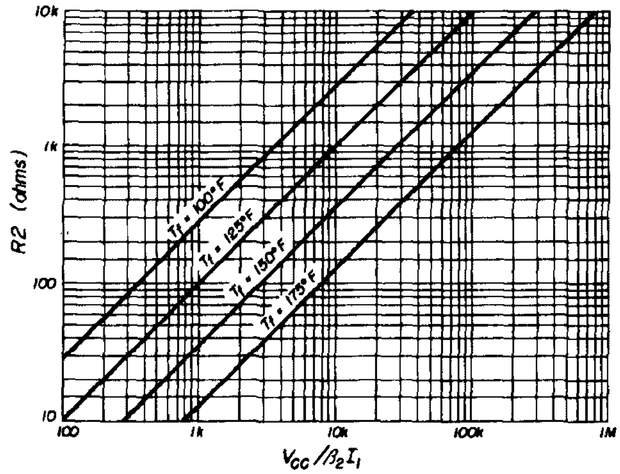



**fig. 2. Value of R2 as a function of Vcc/β2 I1.**

100° F. The measured initial current, at room temperature, was 0.9 ampere; whereas the calculated value was 0.8 ampere. Once the oven had risen to its final temperature, the current required to keep it there was less than 0.1 ampere.

## construction notes

It's important to keep the oven as small as possible. Heat losses depend upon the surface area of the box, so if you increase the size by a factor of two, heat losses will go up by a factor of four. The oven should be insulated with a layer of styrofoam or similar material around it. A ½-inch-thick layer is usually sufficient. Make sure that Q3 is in good thermal contact with the oven wall, since you're relying on thermal conduction to distribute the heat around inside the oven. Put Q1 as close as possible to Q3, preferably in contact with it; otherwise the oven will take a long time to stabilize and may even oscillate in temperature.

**ham radio**

# ssb converter

## for

## 432 MHz

Five watts

of ssb power

with this circuit

outperforms a-m

on the long hauls

Jim Brannin, K6JC, 424 Anson Avenue, Rohnert Park, California 94928

**How would you like to reach** out and work some of those hard-to-get stations on 432 MHz? The answer to extending your DX range is single sideband. Also, if you like to build your own equipment, this mode offers a challenge that's hard to pass up.

The 432-MHz converter described here delivers five watts to the antenna. That's a pretty respectable amount of power at this frequency, considering that the converter was built, for the most part, from "junk-box" components. The only expensive items were the 6939 tubes. The performance of the converter has certainly justified the cost of these tubes. Many stations were worked that just couldn't be contacted with the same power on a-m.

### circuit description

The converter is driven with a six-meter ssb signal; 29 MHz could be used by changing the crystal string to provide 403-MHz output. Obviously, this would require a small amount of coil pruning in the stages. Type 2C51 (5670) tubes and a 6J4, purchased from surplus, make up the crystal string and provide sufficient injection to the 6939 mixer. Two stages operate as doublers and two as grounded-grid amplifiers (**fig. 1**). The 382-MHz signal is link coupled to the mixer grid circuit, because direct coupling between the 6J4 and 6939 tank circuits would require a very long chassis or some sort of double-decking arrangement.

Resistors are used instead of rf chokes to feed the B-plus to the mixer and linear amplifier plate tank circuits. Rf chokes caused spurious radiation and unstable operation in these stages.

All filament, B-plus and bias wiring is on top of the plated epoxy boards. Feedthrough capacitors deliver these voltages to the converter stages. This minimizes stray coupling between stages.

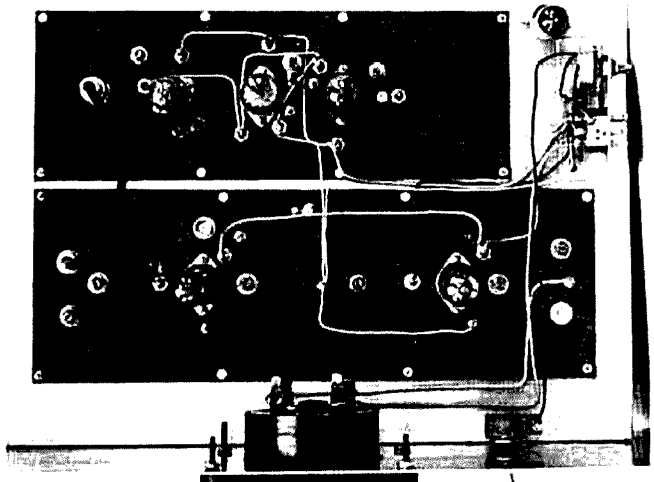
A separate voltage-regulated power supply was necessary to obtain stability in the 95.5-MHz oscillator during modulation. Voltage regulator tubes in the main power supply just wouldn't do the trick.

## construction

The aluminum chassis is 10 x 14 x 3 inches, with two sections cut out to accommodate the two epoxy boards. The cutout for the crystal-strong board is 9-3/4 x 3-1/4 inches; the mixer-amplifier cutout is 11-3/4 x 3-3/4 inches. The boards are 10-1/2 x 3-3/4 inches and 12-1/4 x 3-3/4 inches respectively. Tube sockets, tank circuits, coils, etc., were located for best coupling between stages.

Parts should be placed to obtain the very shortest leads. The input and output circuits of V2, V4A and V5 should be separated by copper shields, extending ap-

**Above-chassis construction of the converter. The oscillator and multiplier stages are on the smaller board to the rear; mixer and power amplifier are in the front.**

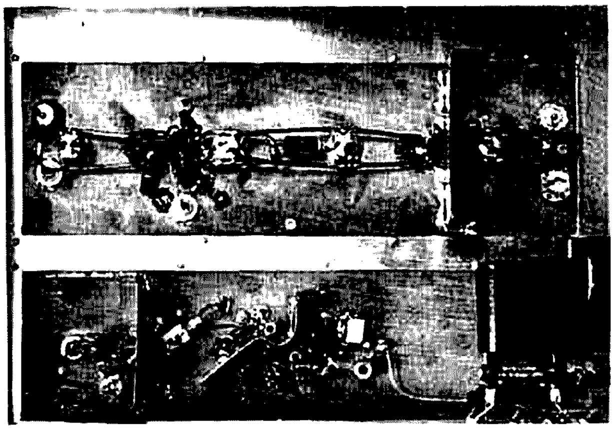


proximately $2\frac{1}{4}$ inches down from the bottom of the boards. The circuit might operate satisfactorily without these shields; however, room should be left so they can be added later if required.

A 3 x 1-3/4-inch aluminum bracket supports the six-prong Jones male receptacle and bias potentiometer as shown in the upper right hand corner of the photo. The OC5 regulator tube in the oscillator power supply is mounted on the edge of the chassis beside this bracket. The standby switch is on the right side of the plate milliammeter on the front panel. The 6.3-V filament transformer, diode rectifier, filter capacitor, etc., are mounted directly beneath the bracket.

The 3/4-wave mixer grid input hairpin, L2, is bent up at 90 degrees as shown in the coil data. This is necessary so that the

**Below-chassis view shows component layout and construction of tuned lines.**



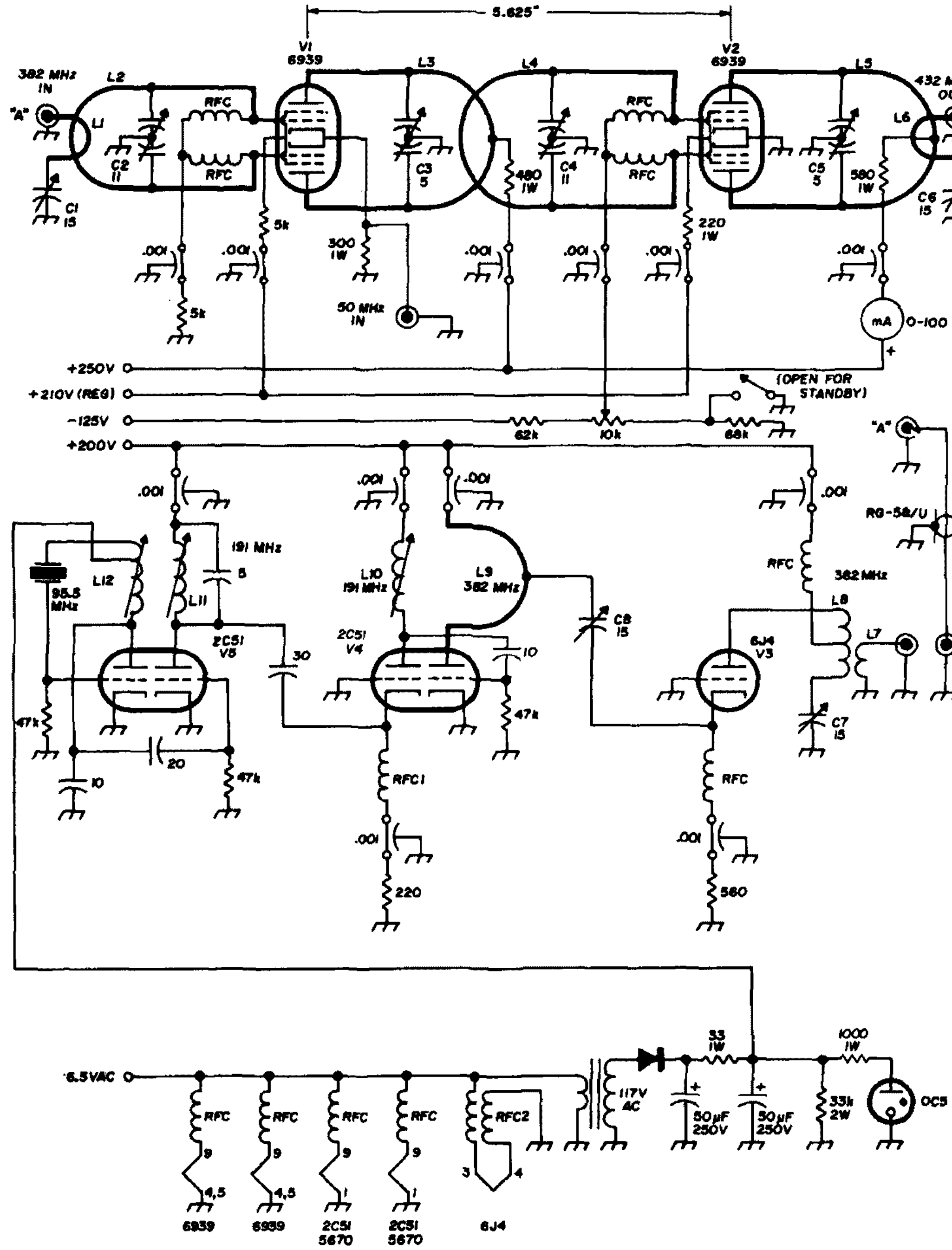
mixer and amplifier stages can be mounted on the $12\frac{1}{4}$-inch epoxy board.

## testing and adjustment

If the coils and loops are made according to the dimensions in the coil and tank-circuit data, little or no pruning should be necessary. A grid-dip oscillator is very helpful. If a grid-dipper is used, the adjacent loop or coil should be wrapped with metal foil, otherwise two resonance indications may be noted unless both circuits are, by chance, tuned to the same frequency.

Tuned-circuit resonance can be checked with a vtvm equipped with an rf probe. Each stage can be checked for maximum drive by coupling fairly closely, either by a two-turn coil or a 20-pF capacitor fastened to the tip of the rf probe. The coupling can be reduced for final adjustment.

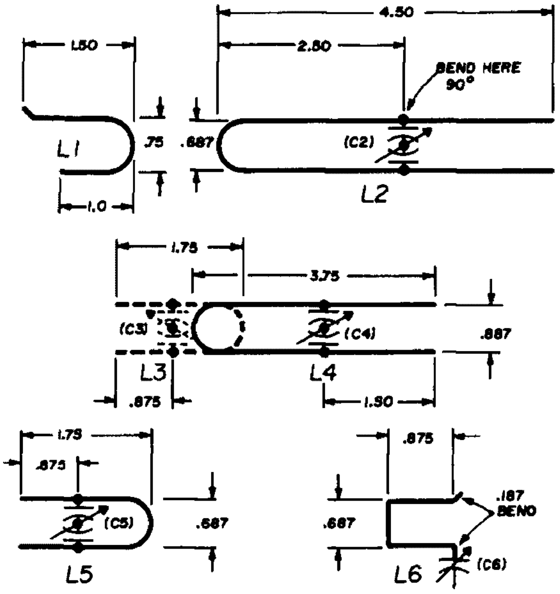
When the crystal string is operating properly, the output at the BNC connector should light a number 47 lamp to about half brilliance. The rf probe can be used




| | |
|---|---|
| C1, C6<br>C7, C8 | 15-pF miniature variable (Johnson 160-107) |
| C2, C4 | 11-pF butterfly (Johnson 160-211) |
| C3, C5 | 5-pF butterfly (Johnson 160-205) |
| L7 | 2 turns no. 18, 1/2" diameter, tightly coupled to L8 |
| L8 | 2 turns no. 18 bare, spacewound, 5/8" diameter, 5/8" long |
| L9 | hairpin loop, 3/4" long, 5/8" wide at ends, C8 connected to center |
| L10, L11 | 3 turns no. 24 on 3/16" slug-tuned coil form, 5/16" long |
| L12 | 6 turns no. 24 on 1/4" slug-tuned coil form, 5/8" long, tapped 2 turns from crystal end |
| RFC1 | 20 turns no. 24 on a 5k, 1W resistor |
| RFC2 | 2 coils; each 8 turns no. 24 bifilar wound on 1 W resistor |

All other rf chokes are Ohmite Z-460 or equivalent.

fig. 1. Schematic diagram of the 432-MHz ssb transmitting converter. Coils L1 through L6 are shown in fig. 2.




NOTE: L5 & L6 OVERLAP 1/2", WITH 1/8" SPACING
L3 & L4 " 5/8", " " "
L2 IS BENT UP AT 90° AND BNC CONNECTOR AND C1 MOUNTED CLOSE TO BEND. L1 IS VERTICALLY MOUNTED TO PARALLEL L2 WITH 1/8" SPACING

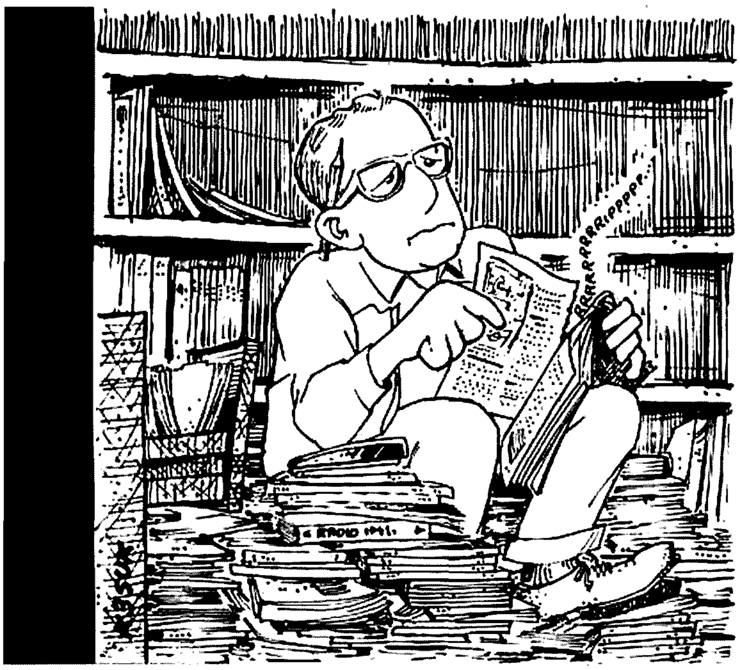
**fig. 2. Construction of the coils for the 432-MHz transmitting converter.**

on the mixer plate tank for preliminary adjustment of input and output circuits of this stage. Plate current in the final will indicate proper adjustment of the 6939 grid loop. Also it's the best indicator for final tuning of all stages in the converter. The final plate current should be approximately 70 milliamperes when the converter is properly tuned and adjusted.

If a thru-line wattmeter is not available, a forward-power indicating device should be installed in the coaxial cable to the antenna. This can be a piece of enamelled magnet wire sewn beneath the coax shield for about three inches, located fairly close to the antenna change-over relay. The end toward the antenna should be connected to a small diode and bypassed; the other end should be terminated with a 50-ohm resistor. A 0-500 microammeter or 0-1 milliammeter can then be connected and operated remotely for convenient viewing.

The converter described here has required no retuning or adjustment for several months, and its performance has been most gratifying.

**ham radio**



# use your magazines

Most of us collect technical magazines, and there they lie unused on the shelf; this article gives some good tips on how to keep them alive

E. L. Foster

**I have finally solved** the problem of what to do about all my old electronics magazines, but it takes courage. It's simple: you take all the new and old copies of the magazines you have, carefully extract the articles that interest you, and throw the rest away! Admittedly, you may change your tastes in future years, but this is better than burying all the articles in a relatively unavailable form in anonymous but attractive rows of magazines. A book in hand is worth two in the bookcase.

## how?

Carefully take the staples out of magazines like **QST**, **Radio-Electronics**, or **Electronics World**, and tear out noteworthy articles. Make sure you include any "continued-on-page-108" items. Most other electronics magazines don't have staples directly through them, so you can tear out each article directly. "Perfect-bound" (i.e., glued-together) magazines like **ham radio** should be folded as flat as possible at the relevant page before tearing out.

Once you've torn out an article, staple the pages together at the upper left corner if there's more than one sheet. Carefully *put them into a Manila folder and invent* a suitable title. (The choice of titles is important and will be covered later.) A title such as "Transmitters" is a lot more useful than "Transmitters—complete—80M — 6 tubes—2 diodes," for example.

### why?

It makes sense. Out of the monthly deluge of information you receive, if you need to know something you merely look up the appropriate folder and it's there. My method also has the advantage of forcing you to read the articles, even if briefly. Space is also saved by eliminating chatty columns and ads.

Sometimes I wonder if the one or two articles I get from a magazine are worth the cost of that issue. For the most part, they are. **QST** is worth saving because of its occasional gems, and the ARRL is worth supporting. **Radio Communication** is worthwhile because of items such as "Technical Topics," and because the RSGB is also worth supporting.


### the file cabinet

If you don't have a filing cabinet, the folders can be kept in a cardboard box. The box can be on an *inconspicuous* shelf, because you'll only refer to it when you're going through current magazines or when you need information. Or you can select magazine covers with unusual designs and colors, trim them to make a pleasing arrangement, then glue them to the sides of the box (known to artistic types as a collage). Give the box several coats of clear lacquer. This reinforces the box.

My storage and retrieval method also works well for old issues–and you'll be amazed to discover the many excellent and useful articles on all manner of subjects. The thrill of this discovery has kept me at this task: five magazines to shred every day until they are all done. It's cheaper to reduce your magazines in this manner than to build new bookshelves to accommodate the growing pile of literature. (Some take the easy way out and merely sell the whole lot after X years.)

The system I use is different from that in which you bind back issues of magazines. These techniques[1, 2] are valuable for binding books, but they have no place in a Magazine Shredding Program. They merely codify the situation but don't solve it. It's rather unproductive to bind a book permanently, 90 percent of which is useless and takes up valuable space, and 10 percent of which you'll never bother to consult because you don't know where it is.

### the system

What do you do if you have two good articles that have been published back-to-back? First decide which is the most interesting. This is the article that should remain complete. There are several cases of this situation, in which:

**1. The less-interesting article starts on the back of the good one.** Write the title and subject briefly at the top of the second page of the amputated article, indicating where the rest of it is filed (**fig. 1**).

**2. An article ends on the back of the good one.** Write on the bottom of the last available page of the amputated article, "continued on..." or "see... file" (fig. 2).

**3. A good article is sandwiched between two others.** Make out a full-sized sheet of paper showing title, subject, and where the two halves may be found. Then file the paper instead of the article.

cuits" in **Radio Electronics**. The hard way to find these articles is to make a separate sheet of paper for each subject and file it in a "collections" file. Sometimes this requires as many as fifteen separate listings for one month's worth of "Technical Topics" and could be more work than you'd want.

Nothing is worse than a good filing sys-

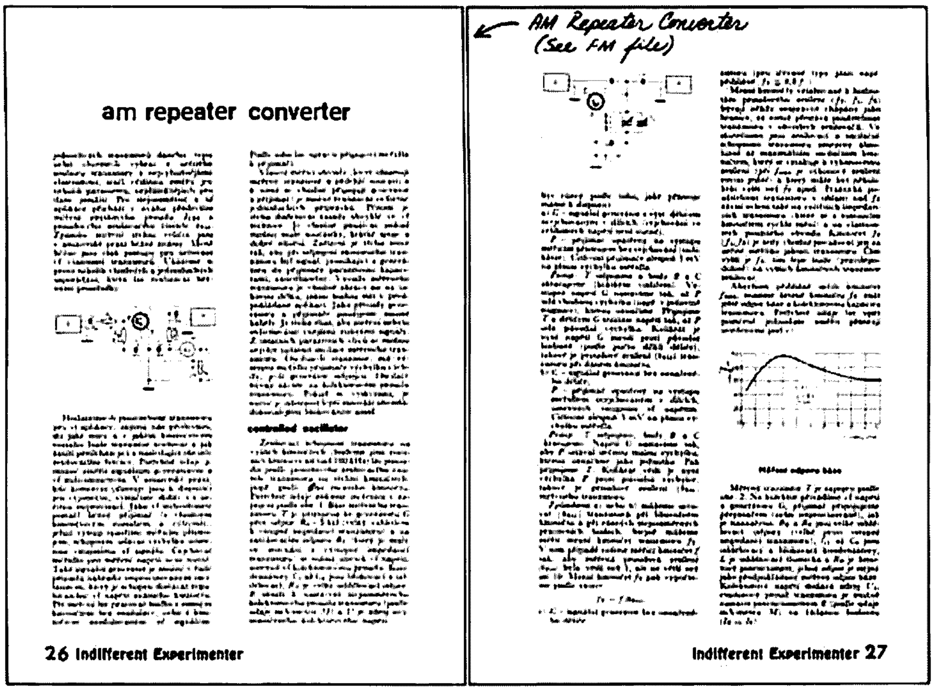



**fig. 1. First two pages of less interesting article. Interesting article, "Advantages of FM Over SSB," ends on page 25. This more important article is filed under "Modulation, FM," but a note of it is entered under "SSB Controversy."**

**4. Two good articles are to remain together.** Write the title of the second article on the front page of the first, and file a descriptive sheet of paper as in 3 above.

**5. Rare publications that don't print name and month of issue on each page.** Examples: **Break-in** (from New Zealand) and **Autocall**. Use abbreviations such as BI 7/67 or AC 7/67.

**6. A single article that contains a wide variety of good designs.** Examples are G3VA's "Technical Topics," which appears frequently in **Radio Communication**; "Hints & Kinks" in **QST**; "Noteworthy Cir-

tem that's not followed, so an alternative must be sought. One way is to file the relevant articles in the "Collections" file for six months. However, you should read them thoroughly each month and make special note of points you may need in the future. Then go back over them, and pick out items according to subject.

You may have the same fifteen or twenty pieces of paper to file under different headings, but each piece will refer to several articles in which a given point is mentioned. For example: one sheet of paper labelled "Transistorized Transmitters" could refer to specific subarticles in four or

five issues of "Technical Topics." I dwell on this problem because it can be frustrating but it is important enough to devise methods for solving it.

## classifying the data

In general, file an article according to the subject, not necessarily the title. Sometimes authors can be misleading in their effort to be clever; editors should never allow this! A recent article bore the title "'The Thing'—Transistorised (sic), an Experimental Sideband Exciter," but the subjects were principally crystal frequency alteration, a bandspread oscillator, an electronic sweep generator and some very interesting data on the use of ordinary rectifier diodes as varicaps. In the next issue the author did indeed get down to sideband, but you would never know it from this one. In this case, I filed it under "diodes, varicap," but I filled out a separate sheet of paper to refer to the article and filed it under "crystals," and another under "oscillators, sweep." I didn't pay any attention to the ordinary oscillator because they are a dime a dozen in the literature.

## data retrieval

There is a definite temptation to keep some items intact, such as "Miniwatt Digest" (Philips) and the like, but such temptation should be resisted, even if attractive binders are provided—which you may already have. Filing away blocks of articles in binders is only justified if you know what's in them. Always ask yourself, "If I do this, can I get my hands on the information when I need it?" If the answer is yes, any filing system is satisfactory. My system is more practical than most, once you get over the horror of the magazine-shredding process.

For a really effective filing system be sure to save the table of contents from each fractured magazine. A cryptic notation can be written in front of each title to show where it was filed, though this is not usually necessary. The table of contents can help where a title might be ambiguous or misleading, or when you are re-

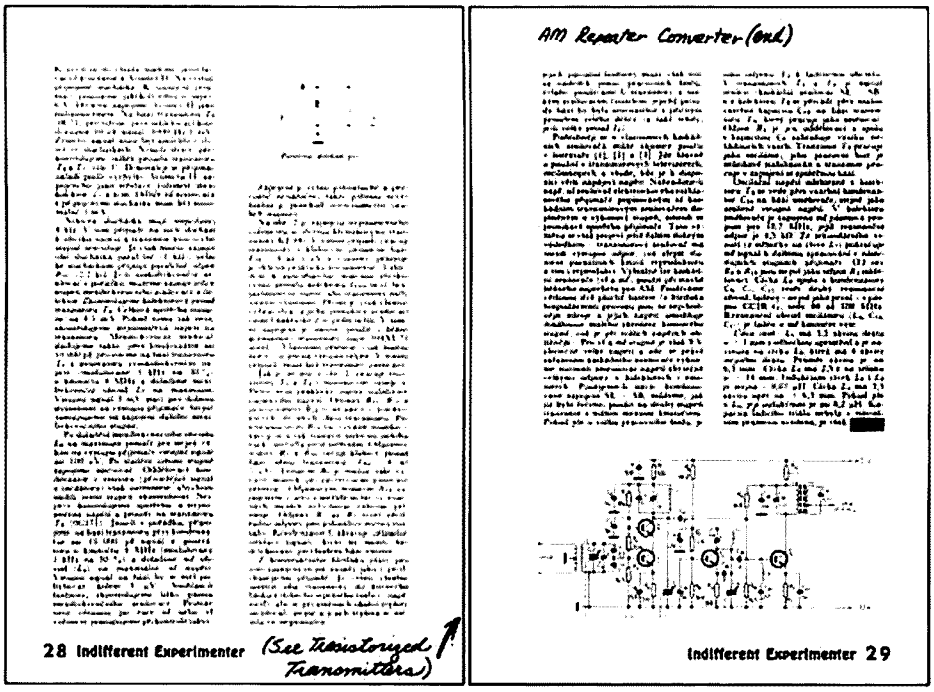



**fig. 2. Last two pages of an uninteresting article. Good article is on page 30.**

ferred to an article by another reference. If I read the **Journal of Indifferent Electronics,** and it refers me cryptically to "**QST,** May 1907, p. 31," the only cure is to dig out the relevant table of contents (which I have filed under "index") and see what it was all about.

The system I describe here is the best of several I've tried. It is crude, but much more effective than any elegant indexing arrangement using file cards or punched cards. The reason is simple: you are more likely to use my system and keep on using it. It's destructive, but what is more useless than a complete set of magazines you never use? Yearly indexes? Nearly useless: the magazine's indexing system is likely to overlook subjects that interest you, but which may be buried in another subject. For example, a good point on transmitter or receiver design may be included in a review of some uninteresting piece of commercial equipment. And let's face it, most commercial "amateur" equipment is uninteresting if it's just a box with knobs.

Advertisements may also be filed by this method if you need them for reference, but you can always get current ads featuring current items. Therefore, don't tear the current issue apart until the next one arrives. Have a definite place to store "magazines to be shredded", and never let the pile get higher than one-half inch!

## what to do about arrl handbooks

An unexpected bonus of the "shred-it-yourself" approach happens in the treatment of the ARRL Radio Amateur Handbooks. As you know, this text appears each year, and well over 50 percent is unchanged from one year to the next. This is where the shredding technique appears at its best. You simply go through the current issue and the previous one simultaneously, page-by-page. This sounds like a lot of work, and it is, but it doesn't take too long.

Since it doesn't usually matter which issue you keep intact, the simplest arrangement is to choose the older issue, or the first one that changed from glossy to dull paper (1962). Then all subsequent issues are torn apart, until about the fifth-year later. In about five years, ARRL handbooks usually change enough to warrant keeping one intact. Then all subsequent ones are referred to it.

## the technique

You have the choice of discarding whole sections, or going through likely sections page-by-page to see where they may have substituted one item for another—usually a whole paragraph. The page-by-page method is simplified considerably by comparing the diagrams; if a section has been altered, a new (or absent) diagram will appear, and will become immediately evident by comparison with the old copy. But if you don't want to go to all this work, you won't be too far wrong if you delete a whole section. You may miss a few interesting items. In this activity a considerable hidden benefit evolves because you are forced to review the entire contents of a text, however casually, and you will be fascinated to discover all manner of useful and interesting things from it.

**1.** From the newer edition you will want to discard completely or in part the chapters on:

Amateur Radio (one is interesting, but is quite sufficient)
Electrical laws and circuits
Vacuum tube principles
Power supply
Amplitude modulation or audio amplifiers & dsb phone
Specialized communication systems
Antennas
Wave propagation
Vhf antennas
Construction practices
Interference with other services
Operating a station—and similar chapters

You must keep one of the editions intact so that you can reread the history of amateur radio if some beginner asks you about it. Save the newest issues and shred only previous ones. The whole subject came up as I contemplated the great stack of ARRL handbooks that had reached the end of

available shelf space. I had purchased one every couple of years on principle, even though I hadn't even opened some of them.

2. You'll want to preserve the construction sections of the following chapters and discard the theory sections, if any. Always save the first page of each chapter, at least, to make it easier to locate:

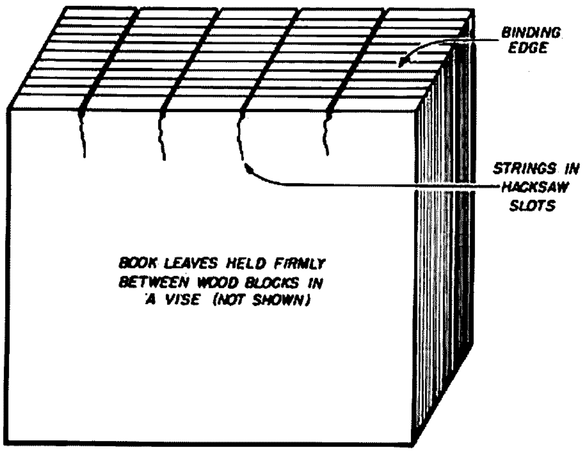



**fig. 3. Simplified bookbinding. It's wise to put the wooden blocks as close as possible to the binding edge. Use waxed paper between the wooden blocks and the book.**

High-frequency receivers
High-frequency transmitters
Keying and break-in
Speech amplifier and modulators
Suppressed-carrier or single-sideband phone
Transmission lines
Vhf receivers and transceivers
VHF transmitters
Mobile equipment
Measurements
Assembling a station
Vacuum tubes and semiconductors; keep the whole chapter because they have a tendency to delete useful older types.
Index: keep all

Some years change more than others. For example, as I recall, there were only about a half-dozen pages in 1964 different from 1963; but 1967 was quite different from 1964—115 pages or 15 percent of the total.

## removing handbook pages

The method of extracting pages may depend on the number to be removed at any one time and on the binding. In older editions the binding was firmer, reinforced by string, and pages will be more difficult to remove than from the newer ones. You have the choice of using a razor blade or carefully tearing a page out from its binding. The tearing method is more practical for whole groups of pages or a whole chapter. Lay a steel rule along the gutter and tear gently.

You should be reasonably careful when extracting pages or sections, because if the alteration to an issue is not too drastic, the decimated issue can be given to a beginner, who will be able to make good use of the theory portions. For this reason, I suggest that when you remove something from an edition, you note carefully on the margin of the previous page the name of the item you've removed; if the recipient is actually interested, he can always dig it up from other sources, e.g., you.

The extracted sections will be added to the reference issue (the oldest one). The best way to do this is to insert relevant material at the end of the chapter related to its subject, in the reference edition. I admit this will result in a somewhat looseleaf arrangement, but it shouldn't matter too much if you're careful. You may not look at it too often anyhow, and when you do, you need only exercise some care in keeping the right pieces together. The problem of increased bulk will be met by the fact that you will have been able to discard the whole advertisement section.

When the requisite editions have been removed in this manner, you can solve the looseleaf problem neatly by binding them together. You can take it to a professional bookbinder and have him do it; the cost won't be much. But it's not difficult to do it yourself if you follow the methods suggested below.

## binding the handbooks

The binding method I prefer is simple

and adequate. (Other material on bookbinding appears in reference 3.) First carefully line up all loose sheets with the fixed ones. Then line up the left-hand edges. If you have a bookbinding shop in your town, the bookbinder will cut the edges in his power-driven cutter. You might also consult your local librarian about this service.

With the edges held together in a vise, make about ten V-shaped cuts with a hacksaw at right angles to the binding, about 1/16-inch deep. Then lay a piece of string in each cut, leaving the ends dangling, as shown in **fig. 3**. Cover the whole binding with glue, spreading it evenly, preferably with a finger, but being very careful not to allow any to slop over any edge. When tacky, lay a piece of linen cloth (as from a sheet) on the binding edge and carefully apply glue all over it. Allow to dry for a day or so, then carefully slice away all dangling cloth and string with a razor blade. Glue the cover to the binding, again being careful not to slop any glue onto the edges of the pages. Apply a one-inch strip of paper or cloth to the point where the cover meets the first and last pages of the book, as shown in fig. 4. Place a piece of waxed paper between front cover and first page, and another between back cover and last page. Close it all up, place another book on it for pressure, and allow to dry for another day.

The result will be a neat looking and functional book, and sturdier than most of the ones you buy. The glue should be the special white liquid bookbinding type because it's very adhesive and flexible. If you can't find it, ordinary white wood glue will suffice. The bookbinding glue is rather expensive, but worth buying, because you'll find it useful in repairing other books, particularly where the publisher has economized on the quantity or quality of the binding.

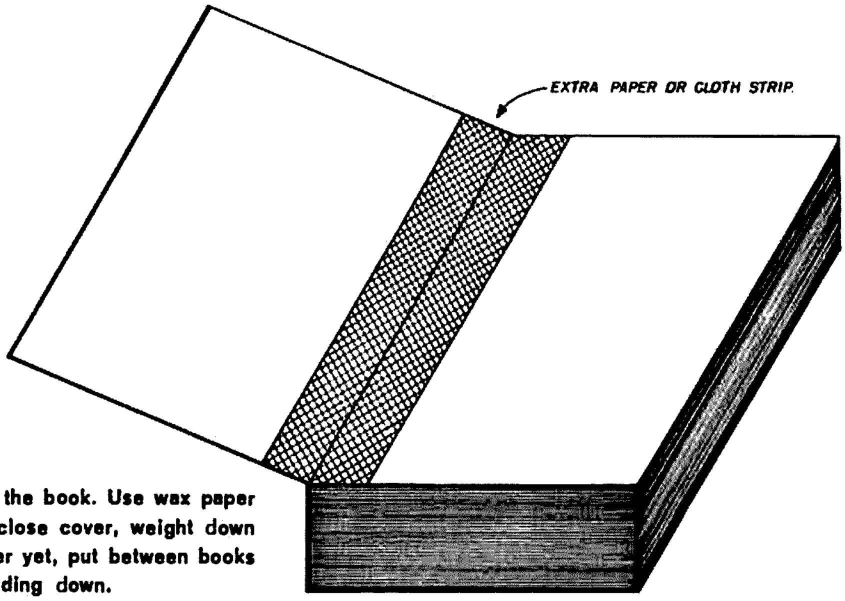



**fig. 4. Putting a cover on the book. Use wax paper between cover and book, close cover, weight down with another book, or better yet, put between books on the bookshelf with binding down.**

## references

1. J. Kirk, W6DEG, "Make the Most of Magazines, and have fun doing it," 73, December, 1966, p. 70.
2. R. Gunther, VK7RG, "Editorial," *Australian EEB*, August, 1967, p. 110.
3. P .W. Waters, G3OJV, "Bookbinding for the Amateur," *RSGB Bulletin*, December, 1966, p. 802.
4. R. A. J. Reynolds, VK7ZAR, "The 1968 ARRL Radio Amateur's Handbook—A Critical Review," *Australian EEB*, December, 1968, p. 130.

**ham radio**

# unidirectional antenna

## for the

## low-frequency bands

End-fire and broadside characteristics are combined in a steerable array using a simple switching circuit

Malcolm M. Bibby, GW3NJY, 2748 Juno Place, Akron, Ohio 44313

**Rotary beams for 40 meters** are too large for the average garden of the European amateur or for most American amateurs' backyards. Many aspiring DX operators who want to work below 7 MHz have tried some form of vertical antenna because of its low vertical radiation angle. Generally what happens is that, after the initial excitement has passed and the DX operator starts the hard work of chasing countries using an omnidirectional antenna, he starts thinking of ways to eliminate noise and interference from locals. To put it mildly, a single vertical antenna isn't too well known for this. The books don't give much help, so if the DXer wants to compete at all he must rely on basic principles and good old amateur ingenuity.

### there is a way

One of the simplest forms of directional antenna systems is the spaced vertical two-element array.[1] The radiation pattern for such a system fed in phase and spaced one-half wavelength apart is shown in **fig. 1**. The beamwidth at the half-power points is 60 degrees.

If one of these elements is fed 180 degrees (one-half wavelengths) out of phase with respect to the other, a horizontal pattern appears as shown in **fig. 2**. The half-

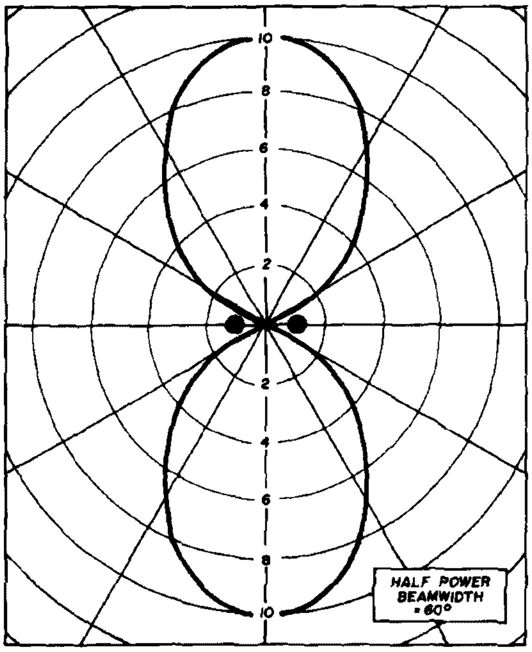
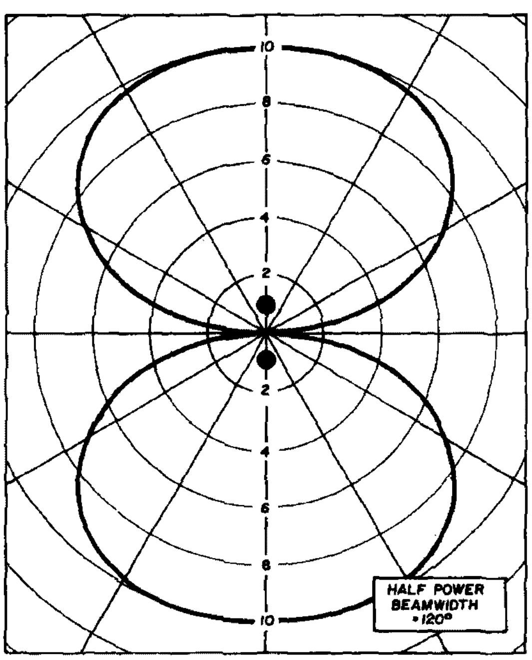



**fig. 1. Horizontal pattern for two V elements spaced one half-wavelength and fed in phase.**

power beamwidth is 120 degrees. Both systems are bidirectional, which is an improvement over the single vertical antenna. Such systems, or a variation of them, have been used by many amateurs.[2,3] However, the problem with either arrange-

**fig. 2. Horizontal pattern for two elements spaced one-half wavelength and fed 180 degrees out of phase.**




ment (broadside or end-fire radiation) is that they are bidirectional and have different bandwidths depending on their phase relationships. This is fine for broadcast service, for which they're primarily used, but broadcast stations don't work DX in the crowded amateur bands.

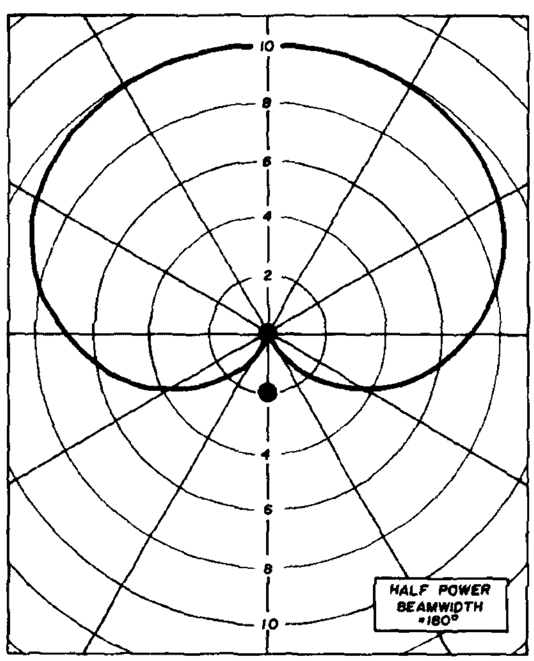
When the spacing between the two elements is reduced to one-quarter wavelength, the phase difference is reduced by 90 degrees. Then the pattern of **fig. 3** results. The beamwith is 180 degrees, but a *unidirectional pattern* has been obtained that can be switched to cover two directions. The beamwidth is as broad as *the side of a barn, though. The question is,* can anything be done to improve it using simple techniques? Yes, indeed. Read on.

## a solution

By using four vertical elements and the principles of the systems whose patterns are shown in **figs. 1** through **3**, a unidirectional system can be obtained having a half-power beamwidth of 88 degrees. It can be made steerable in four directions by simple switching methods.

Consider the four vertical elements of **fig. 4**. Elements 1 and 3 are spaced one-

**fig. 3. Horizontal pattern for two elements spaced one-quarter wavelength and fed 90 degrees out of phase.**



half wavelength. Element 3's phase leads that of element 1 by 180 degrees. Both produce the horizontal patterns of **fig. 5A**. Elements 2 and 4, also spaced one-half wavelength, have the horizontal pattern shown in **fig. 5B**. So far it seems nothing has been gained. However, when elements 2 and 4 are fed with their phases advanced by 90 degrees with respect to element 1, the two lobes in the upper part of **figs. 5A** and **5B** will add, while the two

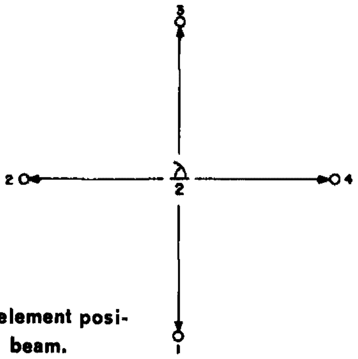



**fig. 4. The four element positions used in the beam.**

lower lobes will cancel. Thus the pattern of **fig. 6** is obtained.

The beamwidth at the half-power points, while not as narrow as a three-element Yagi or quad working under optimum conditions, is nevertheless a respectable improvement over that of a single vertical element.

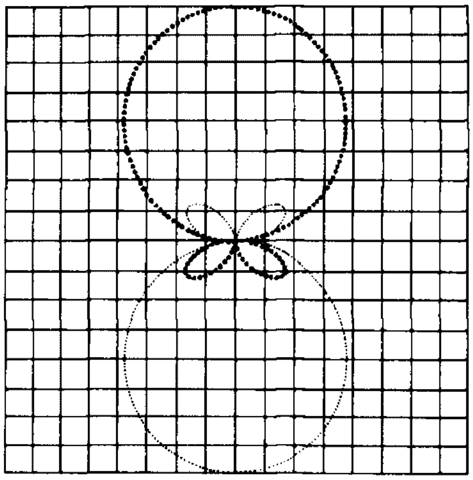
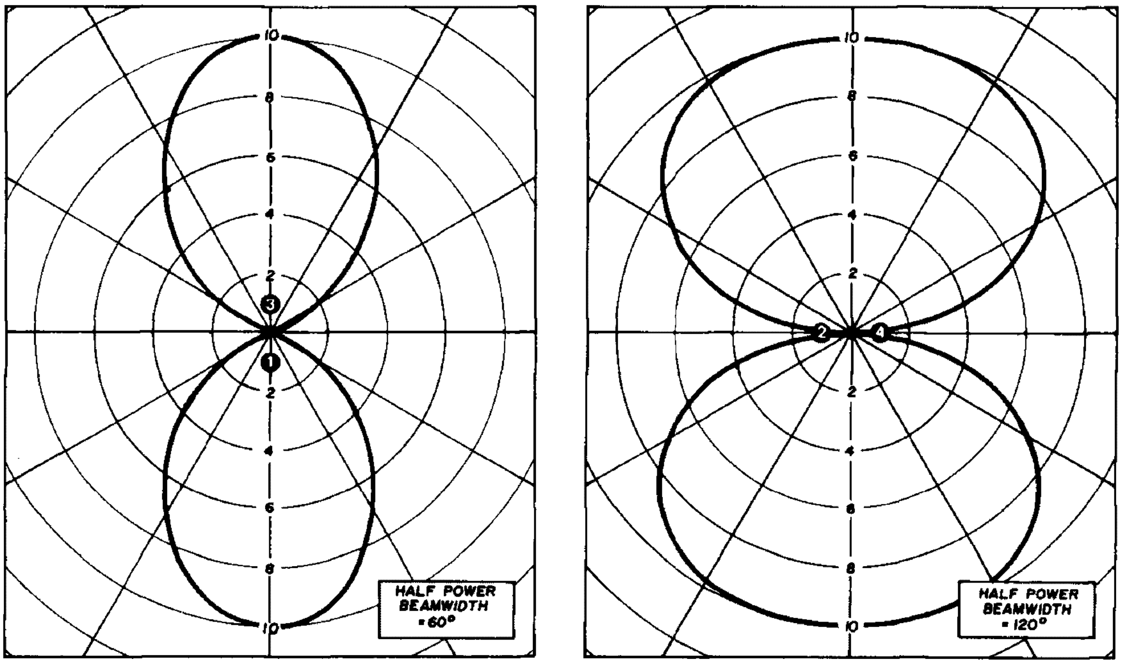


**fig. 6. Computer-derived oscilloscope display of the horizontal polar patterns of the beam. The display of the less-dense dots was achieved by interchanging the feed lines to elements 1 and 3.**

## practical considerations

Like many antenna systems, this one is frequency sensitive, and dimensions should be chosen for the part of the band of greatest interest. As the operating frequency moves away from the design frequency, the small side lobes will decrease but the main lobe will broaden very rapidly—for frequency deviations of four percent or more.

**fig. 5. Horizontal patterns produced by elements 1 and 3 (A) and 2 and 4 (B).**



A further point to remember is that in the theory used to design the beam the assumption is that each vertical element will not only have an identical horizontal radiation pattern (circular in this case) but will also have an identical radiation pattern in the vertical plane. Thus the four elements should be **identical in size** and placed over similar ground systems. For low-angle radiation the ground systems should be as extensive as possible. At the base of each element a rod should be driven into the ground at least six feet with all the radials electrically connected to the ground rod. The radials, which can be any length (the longer the better) ideally should be equally spaced about the base of the vertical with a good electrical connection to the ground rod. The elements don't have to be exactly one-quarter wavelength high, but whatever their height, they must be the same length.

## impedance matching

The elements will have a reactive as well as a resistive component. If the element length is shorter than a quarter wavelength the reactance will be capacitive; if longer, inductive.

A simple network for matching low-impedance transmission lines to a wide range of impedance is the L network. The two basic circuits are shown in **fig. 7**. Both use one capacitor and one inductor of nearly equal reactance. The equations are for matching resistive loads only but provide a starting point for matching R ± jX loads:

$$X_1 = \pm \sqrt{Z_0(R - Z_0)}$$

$$X_2 = \pm R \sqrt{\frac{Z_0}{R - Z_0}}$$

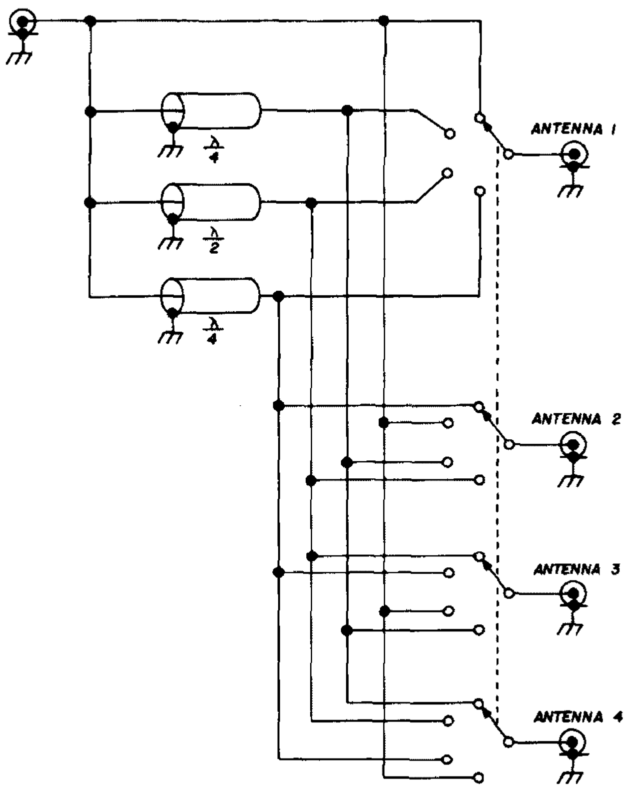



**fig. 8. The wiring diagram for the phasing switch.**

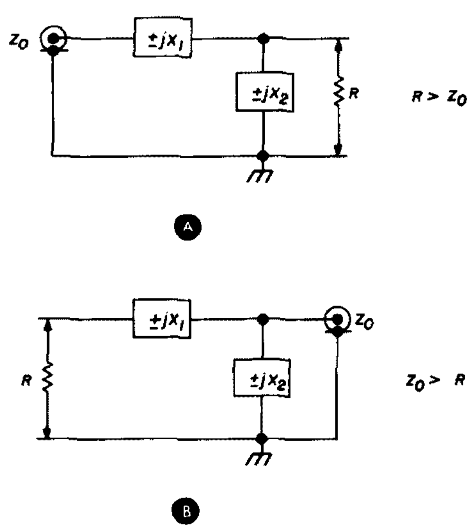



**fig. 7. The L networks, which can be used for matching the coaxial cable to each element.**

Either $X_1$ or $X_2$ can be the inductive reactance, so a maximum of four variations is possible. By connecting a reflected-power indicator in the coax line to each element, each circuit variation can be tried until no reflected power is indicated. The antenna impedance will then be matched, and a weatherproof container can be constructed and placed at the base of each element to house the matching networks.

The impedance match of each element should be checked, because there will be some element interaction. When each element is matched to a coaxial cable all elements should be connected through

equal lengths of cable (the specific length is immaterial) to a control box placed at the center of the system. This box should contain a rugged 4-pole, 4-position switch; preferably remotely controlled. The switch wiring diagram is shown in fig. 8.

After passing through the switch and the phasing cables (working backwards from the antenna), the four lines are joined in parallel. This has the effect of presenting an impedance of 12.5 ohms (for 50-ohm cable) to be fed from the single line from the transmitter. An impedance-matching transformer is needed at this point, and the L-network could be used again; however a better system is shown in fig. 9. This is equivalent to two pi-networks back-to-back. It can be installed in the control box and adjusted for a flat line to the transmitter.

Since devising this scheme I've moved from England to Ohio, and at present I have no means for testing the idea. But the principles are sound and the method should provide an incentive for those who

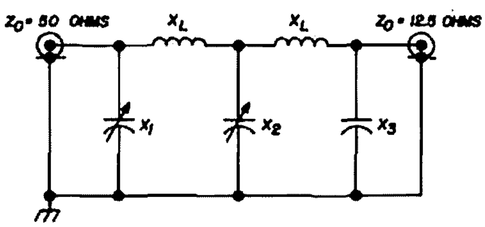



**fig. 9. The 50/125-ohm impedance matching transformer. Typical values would be X1 = 20 ohms, X2 = 65 ohms, X3 = 10 ohms; inductive reactance is 2 times X2 or 130 ohms. Either X1 or X3 can be fixed.**

like to work DX but don't have space or funds to put up a low-frequency rotary beam. The current sunspot maximum has passed, and DX activity will decline on the higher frequencies. I believe this antenna will give you a good chance to compete in the next DX contest.

## references


1. W. D. Stead, VE3DZL, "Twins on Twenty," *QST*, May, 1961, p. 24.
2. I. A. Turner, W9LI, "A Compact Beam for 40 and 20 Meters," *QST*, January, 1954, p. 17.
3. A. D. Mayo, W5DF, "7-Mc Beam for the Small Yard," *QST*, September, 1952, p. 25.


**ham radio**

# counters:

## a solution to the readout problem

Counter decoding is simple and certain with this decade module

Henry S. Knoll, Jr., WAØGOZ, 3392 Ebba Street, White Bear, Minnesota 55110

**Three excellent articles have** recently been published that give the beginner a good basic understanding of integrated circuits and of IC electronic counters.[1,2,3] Each of these counters has one basic complexity however; namely, the readout of the count. The counter in reference 1 reads out in binary, which is satisfactory up to about 10, but nerve-wracking when having to sum 1024, 156, 16, 8 and 1 for example, to count 1305. References 2 and 3 show readout in decades by 10's. This is easiest to read but requires decoding matrices, which are expensive to build and require panel space for

**table 1. BCD counting logic**

| Count | A | B | C | D |
|---|---|---|---|---|
| 0 | 0 | 0 | 0 | 0 |
| 1 | 1 | 0 | 0 | 0 |
| 2 | 0 | 1 | 0 | 0 |
| 3 | 1 | 1 | 0 | 0 |
| 4 | 0 | 0 | 1 | 0 |
| 5 | 1 | 0 | 1 | 0 |
| 6 | 0 | 1 | 1 | 0 |
| 7 | 1 | 1 | 1 | 0 |
| 8 | 0 | 0 | 0 | 1 |
| 9 | 1 | 0 | 0 | 1 |
| 10 | 0 | 0 | 0 | 0 |

ten light bulbs per decade.

A simpler solution is a readout in what is called 1, 2, 4, 8 binary-coded decimal (BCD). With this scheme, which is not new, readout is in decades, but counting within the decade is by binary. Since the highest number count per decade is 9, the binary is easy to decode mentally. This system requires fewer components and less space than straight decimal, because only four light bulbs per decade are required.

the flip-flops are the preset terminals for each unit. These terminals are tied together. When a positive voltage is applied, all of the NOT true outputs go to a high state, and all light bulbs illuminate. The first pulse into the decade will turn all the lights off. This results in a counting error of one, which isn't important at high counts. This can be eliminated if the gating logic of reference 1 is used, which gives an extra pulse. This pulse can be used to clear the preset condition.

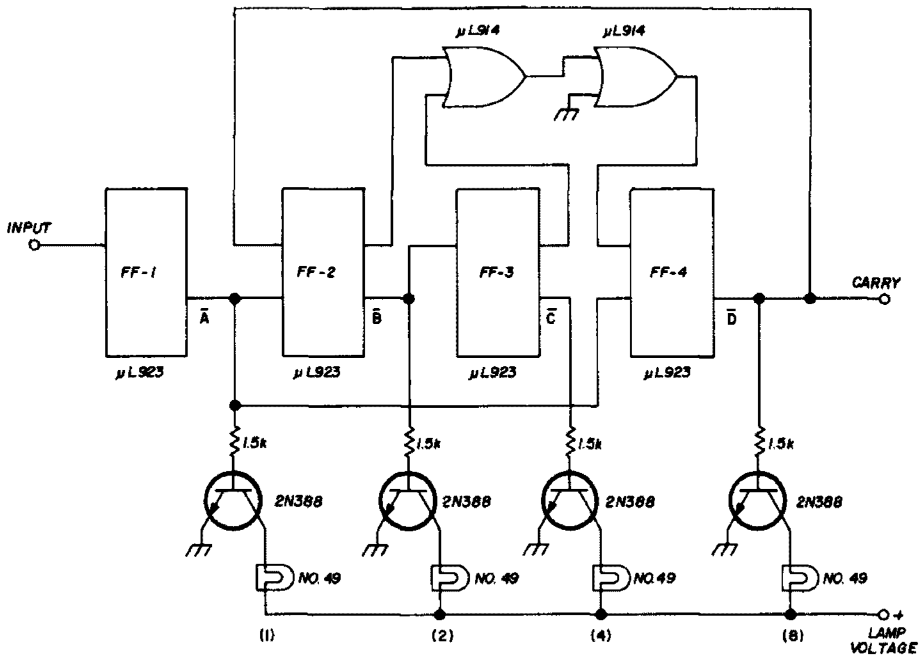



fig. 1. Schematic of the 1, 2, 4, 8 BCD counting decade and lamp drivers.

The counting logic for 1, 2, 4, 8 BCD is shown in **table 1.** A one indicates that the light is on; a zero indicates the light is off. Each decade counts in binary up to 9. The tenth input pulse sets the decade to zero again and passes a carry pulse to the next decade.

The circuit of **fig. 1** shows the simplicity of the counter, and a printed circuit (**fig. 2**) makes the wiring even simpler.

## operation

Referring to **fig. 1**, terminals marked p on

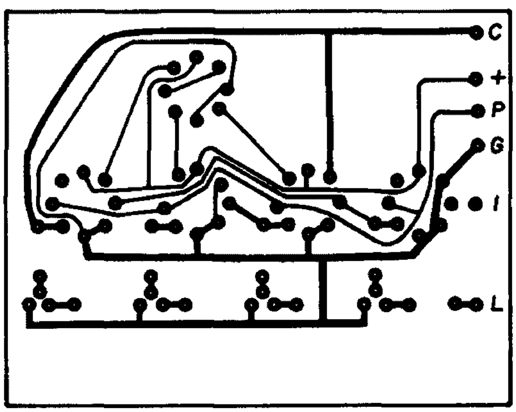
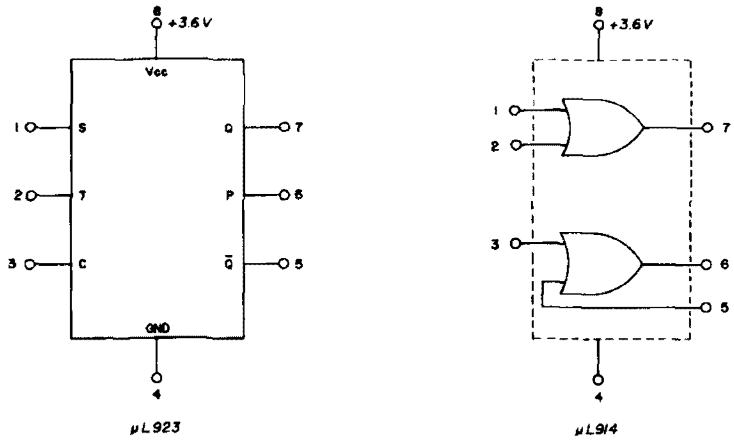



fig. 2. Printed circuit board layout.

fig. 3. Base diagrams for the $\mu$L923 and $\mu$L914, top view.




Numbers in parentheses next to the light bulbs in **fig. 1** indicate the value of that bulb in counting. For example, if the first and last bulbs are on, the count is nine.

## construction notes

All flip-flops are Fairchild $\mu$L923, all transistors are 2N388, all light bulbs are number 49, and all resistors are 1.5 kilohms, 1/4 watt. For those who have never used IC's before, base diagrams are always shown as a top view. The diagrams and operating voltages are shown in **fig. 3.**

Don't expect these flip-flops to work reliably from a battery and switch-closure arrangement. They won't even count with sine-wave input without preconditioning. A $\mu$L914, wired as shown in **fig. 4,** will drive them. Wiring is shown for either switch closure or electrical input.

To make a BCD counter, use all the reference 1 or 2 circuitry except for the counting decades, and add in their place the ones shown here.

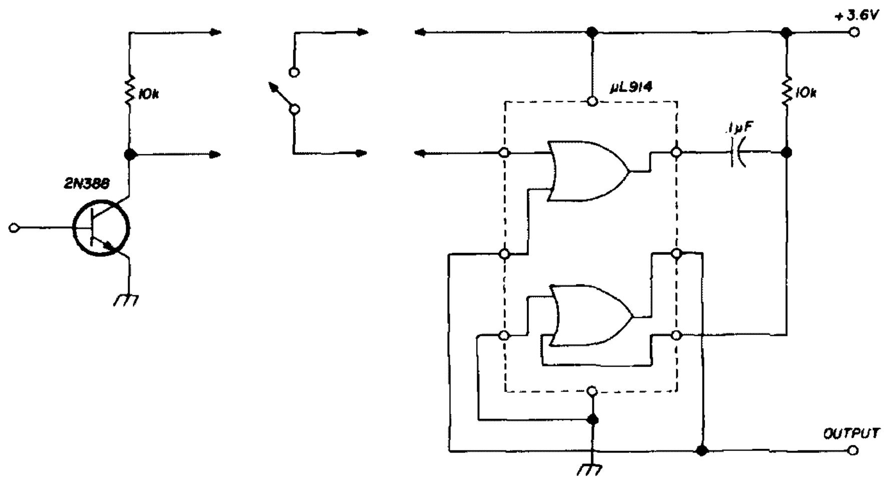



fig. 4. A method of signal preconditioning using $\mu$L914's.

### references


1. R. Suding, W8NSO/WØLMD, "Cheap and Easy Frequency Counter," *73 Magazine*, November 1967, p. 6.
2. G. Jones, W1PLJ, "An IC Electronic Counter," *73*, February 1968, p. 6.
3. D. Lancaster, "Low Cost Counting Unit," *Popular Electronics*, February 1968, p. 27.


**ham radio**

# the ham notebook

## antenna impedance transformer for receivers

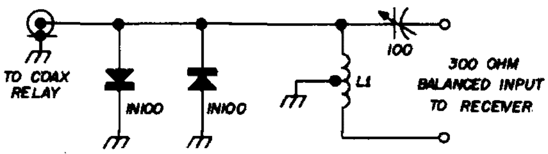



**fig. 1. Receiver antenna coupler. L1 is 10 turns no. 12 enameled, tapped at the center.**

Amateurs spend a great deal of time and effort to obtain an impedance match between transmitter and antenna. The reason is obvious, of course. They want maximum power transfer with minimum loss; besides, it's just plain good engineering practice.

While visiting several local stations, I was impressed with the fact that many hams neglect the other end of the circuit: matching their antenna to the input impedance of their receiver. Receiver manufacturers have no way of knowing what type of antenna will be used, so a compromise is made in the receiver input circuit. The receiver is usually designed for a balanced 300-ohm load.

If you run a length of 50-ohm coax from your antenna relay to this type of input circuit, an impedance mismatch of 6:1 will result. A little figuring will show that the input signal, with this mismatch, will result in a substantial loss.

The little gadget shown in **fig. 1,** mounted on a small phenolic board and attached to your receiver input terminals, will increase your receiver sensitivity by as much as 6 dB (voltage ratio of 2).

If the coil and capacitor combination has a Q around 10 or so, you can adjust the trimmer for maximum response and the circuit will work without retuning on each band. If the Q is higher, of course, you'll have to readjust the trimmer for greater frequency excursions.

A note of caution: Most coax antenna relays have quite high isolation (up to 40 dB) between transmitting and receiving positions. However, with this transformer in the circuit of a receiver having an fet front end, the additional increase in input signal could damage the fet; relay isolation notwithstanding. A couple of diodes, placed as shown in **fig. 1** will protect the fet from overload.

**Alf Wilson, W6NIF**

## mounting bnc connectors

Whenever I mount a BNC connector on a panel, I end up scratching the panel as I tighten down the connector's retaining nut. The answer to this problem lies in a simple homemade tool. I cut the tip off an old screwdriver and epoxied a BNC female fitting to the shaft with metal epoxy. Now, when I install a BNC panel connector, I use the modified screwdriver to hold the connector while I tighten the nut—no more frayed tempers or scratched panels.

**Elliott Kanter, W9KXJ**

## galaxy feedback

When using the Galaxy V MK2 transceiver it's not unusual to have trouble with rf feedback. This is particularly annoying on 10 and 15 meters.

In most cases the problem can be traced to a gassy 12BZ6 rf stage. In this transceiver the tube is operating fairly close to the maximum design limits, and as a result it has a tendency to become prematurely gassy. The useful life of the 12BZ6 can be increased by reducing the plate voltage by 40 volts; according to Galaxy Electronics, this does not alter the published specifications of the receiver.

Plate voltage on the 12BZ6 is reduced by simply replacing resistor R90 (220 ohm, 1 watt) with a 5600 ohm, 1 watt unit. Resistor R90 is located on the Vector extension of the 6GK7 driver and is very easy to change.

**Jim Willis, WA5TFK**

## improving alc response in the SB-400 and SB-401

When using the Heath SB-400 or SB-401, there's no rf output for several seconds when switching out of the spot mode. The slow release time of the alc circuit prevents the ssb signal from reaching succeeding stages for several seconds. This seems like an eternity during contest operation.

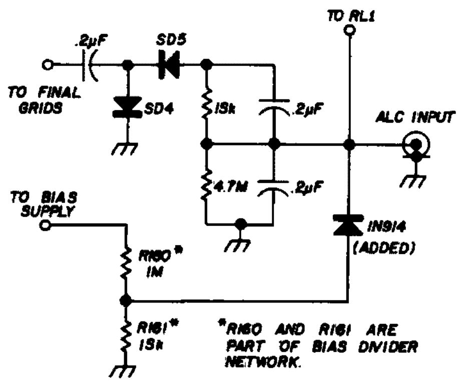



**fig. 2. Adding a diode to improve alc response in the SB-400/401. The added diode should have near-infinity back resistance.**

This problem is caused by a voltage spike passed to the alc network when switching out of the spot mode. The final amplifier bias changes suddenly from -100 V to -50 V, which creates a 50-volt pulse. This pulse must bleed off the alc line before the transmitter can operate.

I solved this problem by installing a silicon diode as shown in **fig. 2.** However, a **word of caution is is order.** Adding the diode places it in series with a 4.7 megohm resistor in the voltage divider on the output side of the alc network. If the back resistance of the added diode is too low, a permanent voltage will be put on the alc line. This will decrease transmitter gain.

For example, assuming a back resistance of 47 megohms in the diode, the 3 V bias on this diode will produce 0.3 V on the alc line. Of course, a lower diode back resistance will put a higher permanent voltage on the alc line. So make sure the diode measures near infinity in the backward direction!

With a good diode, alc voltage will be 3 V. This doesn't interfere with alc operation in my particular rig, but I can't guarantee it won't create a problem in other transmitters.

**David Wojcinski, WA9FDQ**

## modular fm receiver

There has been some misunderstanding regarding the Plectron circuit boards described in the fm communications receiver article in the June, 1969 issue. The Plectron part numbers given in the article were not correct, so inquiries made to the company received negative replies. To readers who had their requests for printed-circuit boards turned down, we offer our apologies for the confusion.

If you are interested in purchasing Plectron boards for the fm communications receiver, send your order to Mike Meyer, Plectron Corporation, Overton, Nebraska 68863. High-band rf board number 3872, $12.95; low-band rf board number 3883, $9.00.

# transistor replaces relay

About five years ago I built a TO-type keyer based on Jim Ricks' circuit. The relay that I used originally was not the recommended (and expensive) type and it lasted about three months. I replaced it with another one that was hermetically sealed; a surplus unit that originally cost someone about $55. It lasted about nine months.

So I tried a cheapie—an open-frame relay that I picked up at a surplus store for 29c. Surprisingly, it lasted four years and was still going when I decided to replace it with a transistor.

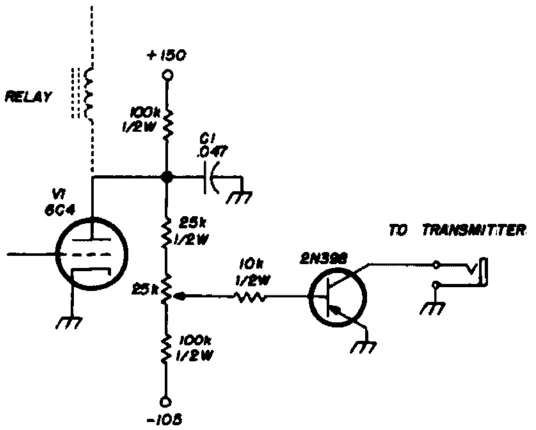



**fig. 3. Solid-state keyer for the W9TO keyer. C1 eliminates any transients generated by the multivibrator.**

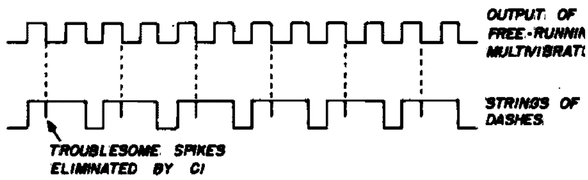
The idea was to do a minimum of reworking on the TO circuitry. **Fig. 3** shows the changes. V1 is the relay driver tube; the relay coil is disconnected from the plate, and the plate is tied to the resistor network shown in the diagram. The 25k potentiometer can be a miniature unit put in some inconspicuous but accessible place; once it's set, it can be forgotten.

The output to the transmitter is the transistor collector. This type of keying can only be applied to a system using grid-block keying where the keyer shorts a negative voltage to ground through a relatively high resistance.

After completing the modification, turn the potentiometer fully counter-clockwise before turning on the power. Recheck your wiring to be sure. Turn on the transmitter and the keyer. When warmed up, the transmitter will come on as though you were holding down the key. Press and hold the dot paddle, and while doing so, slowly turn the pot clockwise until you see that the transmitter is being keyed by the dots. To acquire a feel for the marginal area, turn the pot back and forth and notice the transmitter reaction. Turn the potentiometer clockwise a little past the marginal area—that's it.

The purpose of C1 is to eliminate any transients generated by the primary free-running flip-flop. The transient occurs during the dash period when the free-running multivibrator goes negative. When the relay was used, the transients were not apparent because of the relay's inertia and the short time of the transient. But with a fast transistor, the transient could (and did) appear on the carrier. C1 was selected to provide adequate filtering without materially affecting the dot-space ratio. It did have a slight effect which was compensated by readjusting the keyer weight control.

**fig. 4. Wavetrain when the dash paddle is keyed. The multivibrator transients weren't a problem with the original keying relay because of the inertia of the armature.**



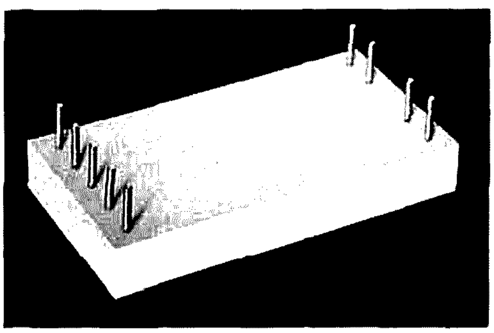

It's a great improvement. No more relay chatter (which the wife appreciates). No more contact bounce, no relay replacements to make, and it's really fast!

A note on the 2N398: The technical data shows that it can hold off 105 volts. The voltage that it's holding off in my Heath Apache is 105 volts. This is really riding it close. However, it's been on the air every night for several months and is still going strong.

**Frank Case, W3NK**

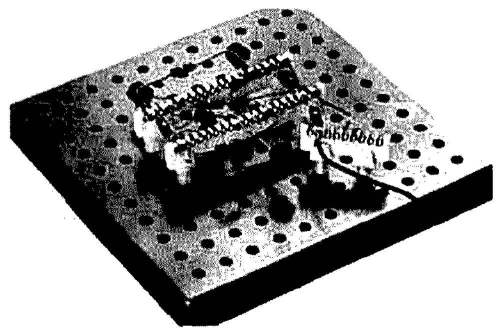
# new products

## hybrid ic amplifier



Bendix Semiconductor is currently offering the first hybrid integrated-circuit device capable of a sustained 15 watts output. The 15-watt class-B audio amplifier exhibits distortion of 1 percent or less while operating into conventional loudspeaker loads. An input signal of 350 mV will drive it to full output; this is compatible with most existing audio preamplifiers. Trimming of idle current and crossover characteristics is provided through external terminals. The BHA-0002 is designed for use in high quality receivers, stereo/hi-fi amplifiers, public address systems, intercoms and music systems. Power gain is 60 dB, frequency response is from 25 Hz to 20 kHz, input impedance is 18 kilohms, efficiency is 60 percent, and noise output relative to 15 watts is minus 70 dB. $9.40 from your local distributor, or write to Bendix Semiconductor Division, South Street, Holmdel, New Jersey 07733.

## breadboard kit



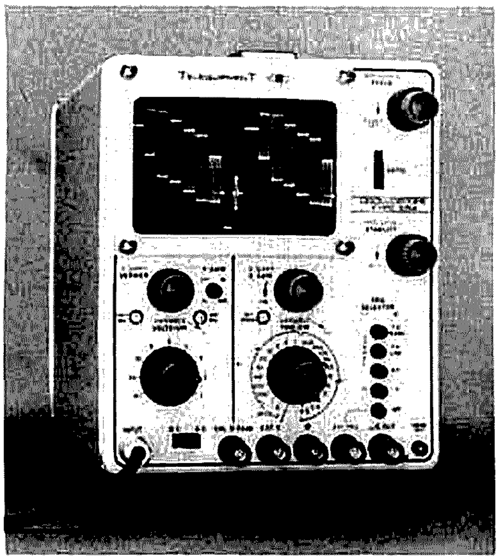
Alco Electronics has recently come out with a design board with pre-punched holes for plug-in Alco terminal strips for breadboarding circuits. The Alcostrips may be quickly inserted and removed from the design board when experimenting with new circuits. The complete circuit-board kit includes 16 Alcostrips in eight different sizes and one design board. $10.95 from your distributor, or write to Robert Laffey, Alco Electronic Products, Inc., Lawrence, Massachusetts 01843.

## tuning diodes

Sophisticated frequency control of uhf circuits is possible with a new series of high-Q tuning diodes from Motorola. The 1N5461-1N5476 series operate over a 30-volt range and are available with a nominal capacitance tolerance of 2 percent (C suffix). This high uniformity is essential where stage-to-stage tracking in tuning is required, but for less stringent requirements 5 percent and 10 percent units (B and A suffixes, respectively) are available.

The nominal capacitance range of these diodes runs from 6.8 to 100 pF. The minimum Q is 600 for the 1N5461A (6.8 pF), and 200 for the 1N5476A (100 pF). The minimum tuning ratio (capacitance at 2 V to capacitance at 30 V) is 2.7. For data sheets with typical design curves, write to Technical Information Center, Motorola Semiconductor Products, Inc., P. O. Box 20924, Phoenix, Arizona 85036.

# telequipment solid-state oscilloscope



The first thing you notice when you unpack the new Telequipment S54 solid-state oscilloscope is it's remarkably compact size. It's only 7-inches wide, 9-inches high and 16½-inches deep including rear CRT protection and panel knobs. Plus it's lightweight—a very attractive package. While the price is more than most of us have been accustomed to paying, it is in line when you consider the excellent specifications and performance. The S54 is built by Telequipment, an English subsidiary of Tektronix. Therefore it's backed by the Tektronix reputation for quality and their excellent field service organization.

The almost 100% transistorized circuit promises long life; power consumption is so low (30 watts) that the cabinet does not get warm even if the instrument is left on all day.

The S54 features triggered sweep, 35 ns rise (per AF book) time, 3 dB bandwidth from dc to 10 MHz and calibrated vertical and horizontal deflection amplifiers. The one item that sets this professional instrument apart from the usual scope is the use of triggered rather than recurrent sweeps. The scope trace is triggered line by line on a selected point of an incoming test signal—depending on the operating mode. This is selected by front-panel pushbuttons which give you a choice of tv frame, tv line or other triggering modes. The two tv modes are convenient for tv service work since the scope easily locks on to the horizontal sync and displays the color burst with no attenuation. In this respect, stability is far superior to lower priced scopes.

The S54 uses an unregulated power supply but the stability is entirely satisfactory. The only tubes besides the CRT are the nuvistors in the vertical input circuit. Vacuum tubes are much more tolerant of inadvertent overvoltages than transistors and can occur if the scope is connected to a strong input signal before the input attenuator is adjusted.

The sweep speed is selected by a calibrated range switch; 22 speeds are available, from as slow as 2 seconds per cm up to 0.2 microseconds per cm. The low speeds might be used by amateurs experimenting with slo-scan tv or ESSA weather pictures. The faster speeds are of interest if you're working with rf or digital circuits.

The price on this versatile new scope is $350. For more information, contact your local Tektronix office or write to Tektronix, Inc., P. O. Box 500, Beaverton, Oregon 97005.

# audio-frequency standard

The Pioneer model 300R is a highly stable solid-state audio-frequency standard that uses a newly developed quartz resonator in a temperature-controlled oven. Up to three standard frequencies may be selected from the front panel, with stability to 0.005%. This instrument Is especially useful for setting up rtty equipment, but may be used for calibrating oscillators, oscilloscopes and bridges. It is also useful for making highly accurate inductance and capacitance measurements. The model 300R functions as either a tone receiver or tone transmitter; the model 300K is a kit version of the model 300 without a cabinet. For more information, write to G. G. Glassmeyer, K6REU, Pioneer Electronics, 738 Pacific Street, San Luis Obispo, California 93401.

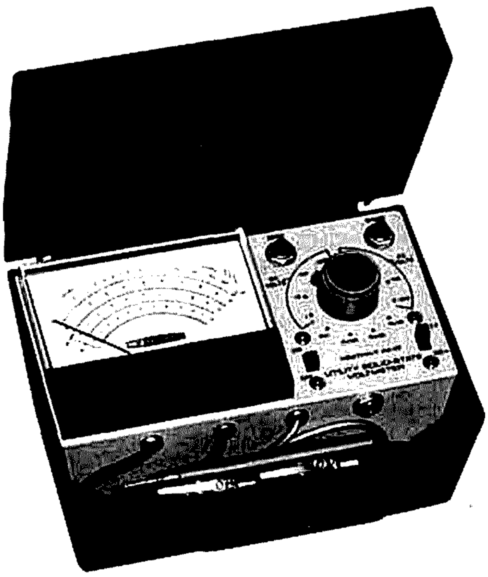
## high-frequency oscilloscope


The new economy-priced Leader LBO-52B oscilloscope features a 3 dB bandwidth from dc to 10 MHz with hybrid circuitry. Its 10 mV/cm sensitivity makes it especially suitable for low-level rf signals in ssb equipment and i-f amplifiers. The wide sweep range has automatic synchronization, and the vertical and horizontal inputs are direct coupled to push-pull amplifiers for distortion-free displays. The units also provide vector pattern display for color tv circuits—this permits viewing patterns at the chroma detector to align the chroma section of the tv set. Priced at $199.00 from Leader Instrument Corporation, 24-20 Jackson Avenue, Long Island City, New York 11101.

## antique tubes

If you're working with old radio gear, but can't find a particular vacuum tube, you might try Arcturus Electronics. They recently acquired nearly 10,000 obsolete tubes, circa 1925-1930, to add to their already considerable inventory of the same hard-to-get types. Listings plus prices of thousands of other items are included in their recently published catalogue, which they will send on request. Write directly to Arcturus Electronics Corporation, 502-22nd Street, Union City, New Jersey 07087.

# solid-state voltmeter



One of the best items to come from Heathkit in quite awhile is the new IM-17 utility solid-state voltmeter. The IM-17 uses one fet and four bipolar transistors plus one diode to provide vtvm performance.

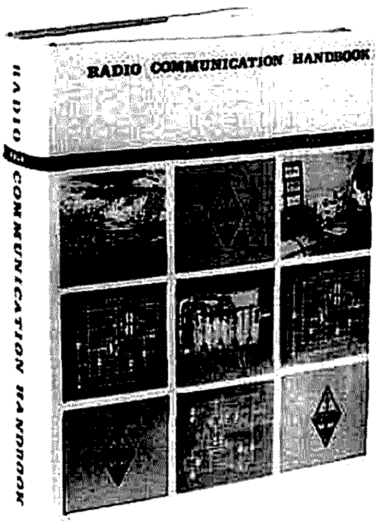
All of the electronic components are on one circuit board; this and the front panel slides into a plastic case which folds over to cover up the whole unit. A handle is molded into the case for carrying.

The ranges of the IM-17 are from zero to 1000 volts ac or dc and zero to 100 megohms. With the fet input circuit this vom acts like a vtvm with a high input resistance so any circuit under test will not be loaded down. The input resistance on the dc ranges is 11 megohms. The dc scale is covered in four positions, as is the ac scale. The resistance ranges are 1, 100, 10k and 1 megohm center scale. There are no current ranges. The only thing not included in the kit are two batteries; one C cell and an 8.4-volt mercury cell. The mercury cell wasn't available at the local radio store but the manual came to the rescue with a recommended substitute.

The IM-17 makes a nice meter for primary use in the shack or as a second one to use on the mobile or field day. If you've been looking for a new meter, here is a vom that acts like a vtvm for only $21.95.

**ham radio**

# partners in excellence




An outstanding technical guide to all phases of amateur radio. In 832 pages 20 complete chapters are devoted to such subjects as single sideband, antennas, mobile equipment, RTTY and much, much more.

This excellent book has received wide acclaim on both sides of the Atlantic and belongs in your library . . . **now.** **$11.95**

**OTHER**
**POPULAR RSGB PUBLICATIONS**

| | |
|---|---|
| Radio Date Reference Book | $2.50 |
| VHF-UHF Manual | 3.75 |
| Amateur Radio Techniques | 2.50 |
| Amateur Radio Circuits Book | 2.00 |
| World at Their Fingertips | 2.50 |

All prices postpaid in USA & Canada


Many thousands of you have become very familiar with the various Radio Society of Great Britain books and handbooks, but very few of you are familiar with their excellent magazine, **Radio Communication.**

This is the oldest and most widely read British amateur radio magazine. Published monthly it provides complete coverage including such popular features as: *Technical Topics, a monthly survey* of the latest ideas and circuits, Four Meters and Down, a rundown of the latest in VHF and UHF and much more.

It includes numerous technical and construction articles in addition to a complete rundown on the month's events in amateur radio. Surely a most interesting addition to your amateur radio activities.

We can now offer this fine magazine to you along with the other advantages of membership in the RSGB (such as use of their outgoing QSL Bureau) for only $6.00 a year.

book division

comtec

Box 592 • Amherst, New Hampshire 03031

"WE ARE THE EXCLUSIVE NORTH AMERICAN DISTRIBUTOR FOR RSGB PUBLICATIONS — DEALER INQUIRIES INVITED"

75 cents

hr

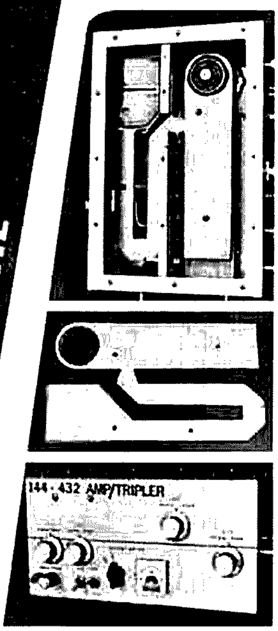
focus
on
communications
technology . . .

# ham radio

magazine

FEBRUARY 1970

144- AND 432- MHz




# DUAL-BAND STRIPLINE AMP/TRIPLER

this month

february 1970
volume 3, number 2

**staff**

James R. Fisk, W1DTY
editor
Nicholas D. Skeer, K1PSR
vhf editor
J. Jay O'Brien, W6GDO
fm editor

A. Norman Into, Jr., W1CCZ
Alfred Wilson, W6NIF
James A. Harvey, WA6IAK
associate editors

Jean Frey
art director
Wayne T. Pierce, K3SUK
cover

T. H. Tenney, Jr., W1NLB
publisher

**offices**

Greenville, New Hampshire 03048
Telephone: 603-878-1441

**ham radio** magazine is
published monthly by
Communications Technology, Inc.,
Greenville, New Hampshire 03048

Subscription rates, world wide:
one year, $6.00, three years, $12.00.
Second class postage
paid at Greenville, N.H. 03048
and at additional mailing offices.

Foreign subscription agents:
United Kingdom:
Radio Society of Great Britain,
35 Doughty Street, London WC1, England.

All European countries:
Eskil Persson, SM5CJD, Frotunagrand 1,
194 00 Upplands Vasby, Sweden.

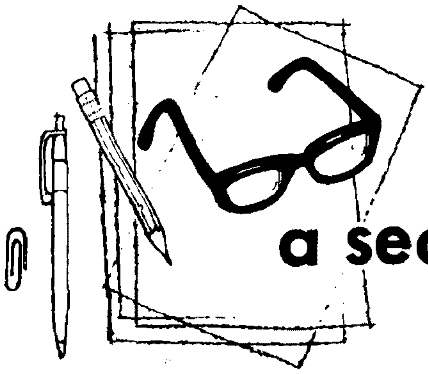

Printed by Capital City Press, Inc. in
Montpelier, Vermont 05602, U.S.A.

**ham radio** is available to the blind
and physically handicapped on magnetic tape
from Science for the Blind,
221 Rock Hill Road, Bala Cynwyd,
Pennsylvania 19004.
Microfilm copies of current
and back issues are available
from University Microfilms,
313 N. First Street, Ann Arbor,
Michigan 48103.

Postmaster: Please send form 3579 to
**ham radio** magazine, Greenville,
New Hampshire
03048.

ham radio magazine

# contents



# a second look

by jim fisk

**One of the most devastating experiences** that could befall a ham is to lose an expensive tower and antenna during a storm. What's even worse is to be forced to dismantle an antenna intallation because of some overlooked restriction pertaining to a local zoning law or building ordinance.

Hams often erect antennas and install electrical wiring at a new location in their eagerness to get on the air, only to receive a "cease and desist" order from local authorities because of some oversight. If you plan to rent, lease, or buy a choice piece of real estate that promises to be the answer to a dream location, it's well to examine the fine print in all legal documents associated with the transaction **before** you sign anything.

In many parts of the country, local building codes contain restrictions against the installation of "radio transmitting apparatus and appurtenances thereto." Such restrictions are often included in lease agreements, title insurance policies, and grant deeds.

Housing developments are becoming a part of the more densely populated areas of the country. Property deeds in these developments often prohibit installation of "structures that conflict with the existing and planned decorative ensemble," or similar phraseology. The language of these documents is often made vague for a purpose: the development planners (and your neighbors) can then ban any structure they might consider to be at variance with local architectural and landscaping motifs. Such structures might include anything from chicken coops to your $500 antenna installation.

Before buying or leasing property, make certain a clause is added to the legal documents that will allow you to install a tower and antenna. It's a good idea to enlist the services of an attorney to help you word the clause, otherwise loopholes may exist that will cause all kinds of problems. The small fee for the attorney's help is well worth the peace of mind after you sign on the dotted line.

As a first approach, consult your local radio club. In most large cities, ham clubs are represented by people who are knowledgeable in this area, or have members who can steer you in the right direction. In California, for example, the Los Angeles Council of Radio Clubs has a service that provides answers to questions from any ham regarding the specific language in a clause to be added to a real estate document. This service is provided gratis by W6QJW and WA6ZCO as a contribution to amateur radio. Many other cities have similar services.

Sometimes electrical and building codes aren't included in real estate documents; nevertheless they're still a part of local law. These ordinances should be thoroughly investigated before you add anything to your property. You might take a tip from the experience of one ham who installed a heavy-duty 220-volt circuit between the power company's service drop and his kilowatt transmitter. Not only did he neglect to obtain a construction permit he also tied several guy wires from his tower to a power company utility pole. He was doing great in a DX contest one day when he was visited by a representative of the local sheriff's office. Sure enough, he was presented with a "cease and desist" order that required him to disassemble his tower and the 220-volt line to his rig. This chap finished the DX contest with an indoor dipole antenna and a healthy respect for local ordinances.

**Jim Fisk, W1DTY**
**editor**

# dual-band stripline amplifier/tripler for 144 and 432 mhz

This state-of-the-art stripline amplifier provides an effective efficient way to get on two popular vhf bands

Richard T. Knadle, Jr., K2RIW, Airborne Instruments Laboratory, Deer Park, Long Island, New York 11729

**Here is an easy way to update** your present 2- to 20-watt two-meter exciter and get on 432 MHz at the same time. Bandswitching is accomplished merely by adjusting the tuning controls on the front panel.

The 500 watt capability on two meters is more than enough for most serious vhf enthusiasts, while the 80 watts available on 432 MHz is a respectable output and will give ample drive for almost any kilowatt final you may build.

The stripline tuned circuits used in this unit yield high Q and result in well-shielded cavities although common aluminum chassis are used throughout. Furthermore, this amplifier/tripler can be reproduced with little difficulty—all cavity parts are made from ordinary flat material with a nibbling tool.

The unusual grid circuit allows independent adjustment of the resistive and reactive

components so that the input to the amplifier can be made to look like 50 ohms. This is an important factor when driving an amplifier with a transistorized 2-meter transmitter since transistors are vulnerable to a high swr load.*

The stripline technique is a modern cavity approach you have probably seen before,[1,2,3,4] and I suspect you will see more of it in the future. Theory of operation is almost identical to coaxial cavities, but striplines are much easier to build, often provide higher Q's, and have some flexibilities which do not exist with conventional cavities. For example, continuously tapered lines are easy to build with stripline, but picture the plight of an amateur searching for continuously tapered pieces of tubing if he were using normal coaxial cavities for an amplifier like this one.

The stripline is placed midway between the ground planes in this amplifier/tripler so equal rf currents flow along both sides of the conductor and full advantage is taken of the stripline technique. In ordinary coaxial cavities, as with stripline, rf currents only flow along the outer surface of the inner conductor; with stripline this conductor can conveniently be made very wide, which keeps losses down.

## operation

The plate circuit for 432, consisting of L1 and C1, is a half-wavelength resonator with a "flapper" variable capacitor for tuning (fig. 1). At the voltage nodal point on L1 where an rf choke for feeding in B+ would normally be connected, a folded half-wavelength line is used instead (L2 and C2, resonant at 144 MHz). If the nodal point on L1 is correctly chosen, the 144-MHz line causes no compromise on 432 MHz.

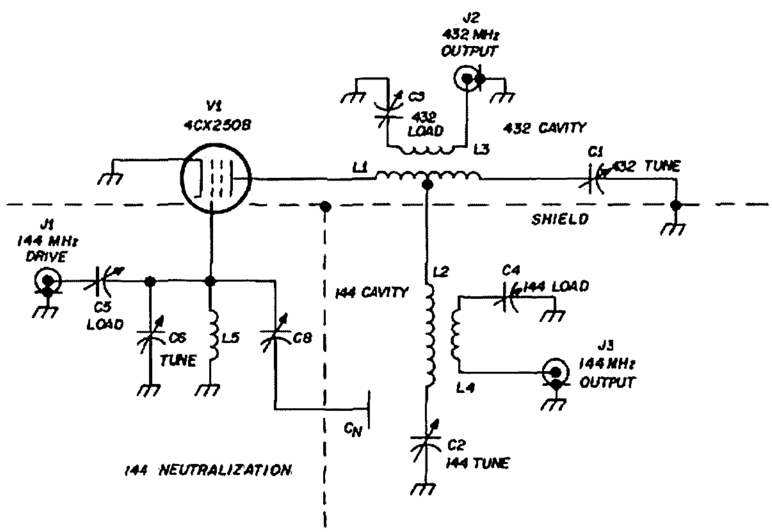
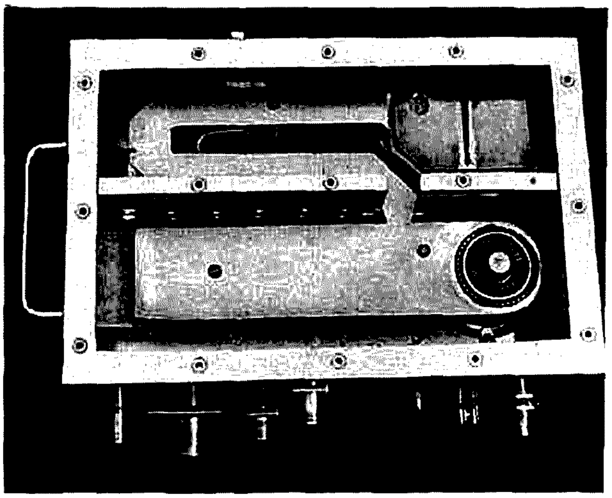



**fig. 1. Simplified schematic diagram of the grid and plate circuits of the 144/432 MHz amplifier/tripler.**

**Inside view shows positioning of copper-clad stripline. Part of L4 is visible inside the bend of the 144-MHz line in the upper left-hand corner; C1 is in the lower left.**



* I drive the amplifier/tripler with a solid-state 20-watt-output 2-meter nbfm transmitter that uses one $30-type transistor. I will be glad to send you a schematic diagram for a self-addressed, stamped envelope.

The components of the 432 cavity act as stray capacitance to the 2-meter cavity, and slightly increases current in the 2-meter section. The inherent high Q of the stripline and the demonstrated efficiency of 75 percent on 2 meters indicates that this is not a problem.

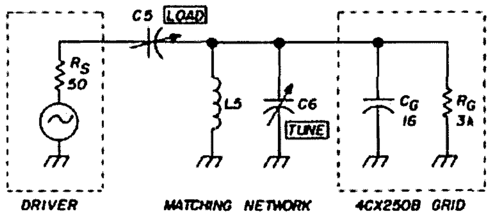
When you desire 432 MHz operation, L1 and C1 are resonated to 432, and the 2-meter cavity is tuned off resonance by minimizing the flapper capacitor C2. The output loop L3 and C3, tuned to 432 MHz, supplies 432 MHz energy to J2.

When two-meter operation is desired, the 432 cavity is tuned off resonance by minimizing C1, and the two-meter cavity is resonated with C2. The output loop L4 and C4, tuned to 144 MHz, supplies output to J3.

The shield between the two cavities and the individually-tuned pick-up loops minimize undesired output frequencies. In addition, the shield breaks the box into two smaller chambers that greatly reduces the possibility of a higher mode resonance; this danger exists whenever two dimensions of a box approach one half wavelength, and this little-known factor has often frustrated amateurs. When it occurs, you can't make the cavity resonate at the correct frequency, find the correct position of the output coupling loop for proper loading, or make push-pull finals share the load equally. When the box itself resonates, it is usually unloaded and has a higher Q than the resonator inside; this causes the box resonance to predominate. All in all, you are safer if you keep boxes small.

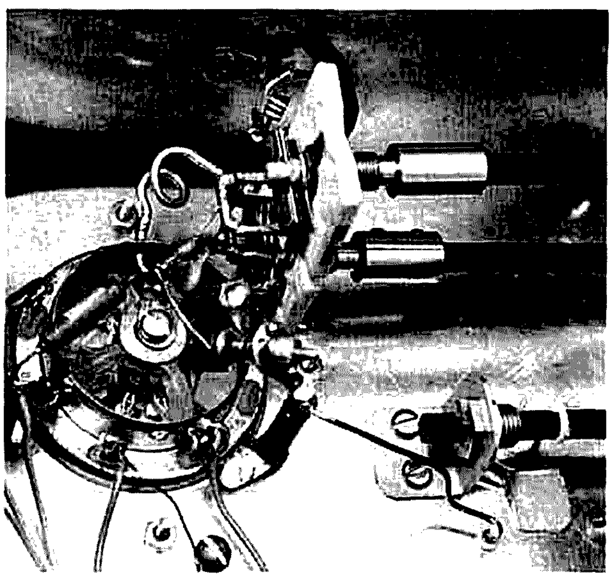
I think it's time that amateurs try something besides link-coupled grid circuits. Motorola has an excellent application note[5] which describes a number of matching circuits and gives values for 50-ohm systems. Although intended for transistor applications, they are applicable to tubes as well. The circuit I used allows the resistive and reactive component to be adjusted over a large range. It is reassuring to put an swr meter between the driver and final and be able to adjust for a 1:1 match at all reasonable drive levels.

During two-meter operation, the grid of a 4CX250B looks like 16 pF in parallel with a 3k resistor. With the greater bias and drive used for tripler operation, this resistance goes up to about 14k. The schematic of the grid circuit is as follows:



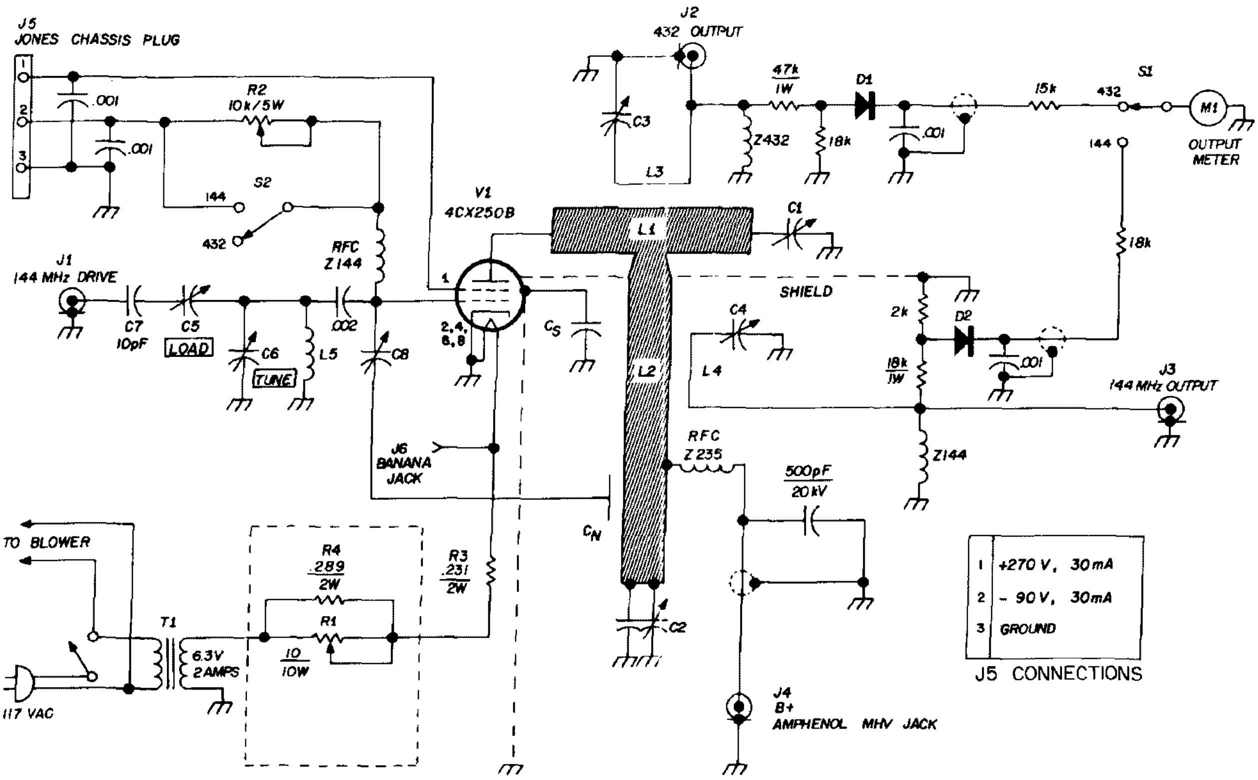

Capacitor C5 controls the transformation ratio (loading) and C6 controls the reactive component (tuning). The values in fig. 2 have enough range for operation on either 144 or 432 MHz. Some stretching or squeezing of L5 may be required to center the tuning of C6. C7 in series with C5, merely sets the range of C5. The resultant grid circuit has a loaded Q of 20 for 144 MHz operation and 75 for 432 MHz operation.

**Grid-circuit layout. Inductor L5 is above socket; C5 and C6 are stacked to the right. The C2 tuning mechanism is in the lower right-hand corner. Stiff wire in lower center is part of neutralizing network.**



## construction

The amplifier/tripler is built in two 8x12x3-inch chassis screwed together top to bottom. The top of the upper chassis is nibbled away, leaving a 3/4-inch wide mounting surface for the cover plate. The tube socket hole is drilled 2-7/8-inches from the




| | | | |
|---|---|---|---|
| **C1, C2** | **Stripline flapper capacitors; see fig. 6** | **D1** | **hot carrier diode (Hewlett-Packard HP-2301)** |
| **C3** | **0.4-8 pF piston (JMC 1802). APC types do not function well here** | **D2** | **Fairchild FD-100** |
| **C4** | **25 pF air variable (APC-25)** | **L1, L2** | **Stripline inductors, see fig. 4** |
| **C5, C6** | **17.5 pF air variable (HF-15)** | **L3, L4** | **Output coupling loops, see fig. 5** |
| **C8** | **0.5-3 pF piston (JFD VC25G)** | **L5** | **1½ turns no. 12, 5/16" diameter, ½" long, with ½" lead length** |
| **$C_S$** | **Screen bypass capacitor, part of the tube socket** | **R3** | **5 feet no. 22 wire, coiled** |
| **$C_N$** | **Neutralizing capacitor, 3/4" diameter brass plate, ½" from L2 in vicinity of C2.** | **R4** | **17.5 feet no. 22 wire, coiled** |
| | | **T1** | **6.3 volt, 3A transformer (UTC S-55)** |

**fig. 2. Complete schematic diagram of the amplifier/tripler. Resistor R3 should be 0.08 ohm.**

rear of the chassis and 2-1/8-inches from the left side. One side of a third 8x12x3-inch chasis was sawed off and shortened slightly to serve as the shield between the two cavities (**fig. 3**).

A less expensive approach would be to bend this shield from a piece of sheet aluminum, but I didn't have a sheet metal brake available. The shield is located 4-1/4-inches from the rear of the chassis. Some of the holes in the shield were positioned to optimize cooling of the two-meter strip-line. Subsequent tests have shown that this was not required. The hottest point on the two-meter line runs 50° C hotter than the cooling air.

I used 6-32 bolts and blind Bendix rivnuts for fastening the top plate. This was done because it was necessary to remove the top cover many times during the stripline trimming process. However, now that the design is completed, self-tapping sheet-metal screws should serve nicely. The stripline is built from a single 6x10-inch piece of 1/16-inch double-sided two ounce copperclad epoxy-fiberglass board (**fig. 4**).

This material was used because it is easy to cut and solder. The conductor thickness is more than adequate since rf current skin depth in copper at 144 and 432 MHz is 0.218 and 0.125 mils respectively. Almost any rigid material with a copper thickness

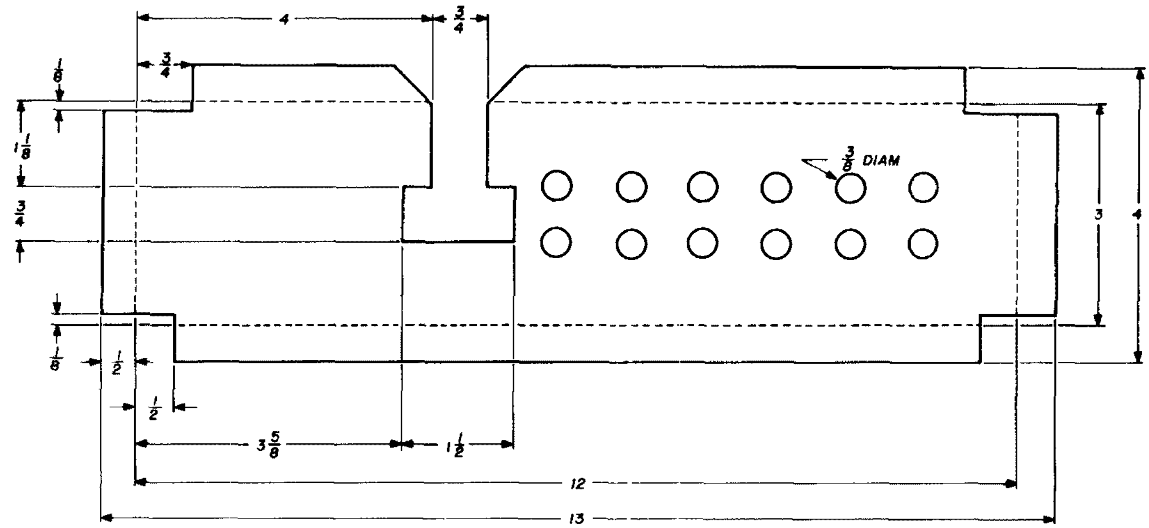



**fig. 3. Layout of the shield used between the 144 MHz and 432 MHz resonant circuits. Shield is located 4¼" from the rear of the amplifier (see photos).**

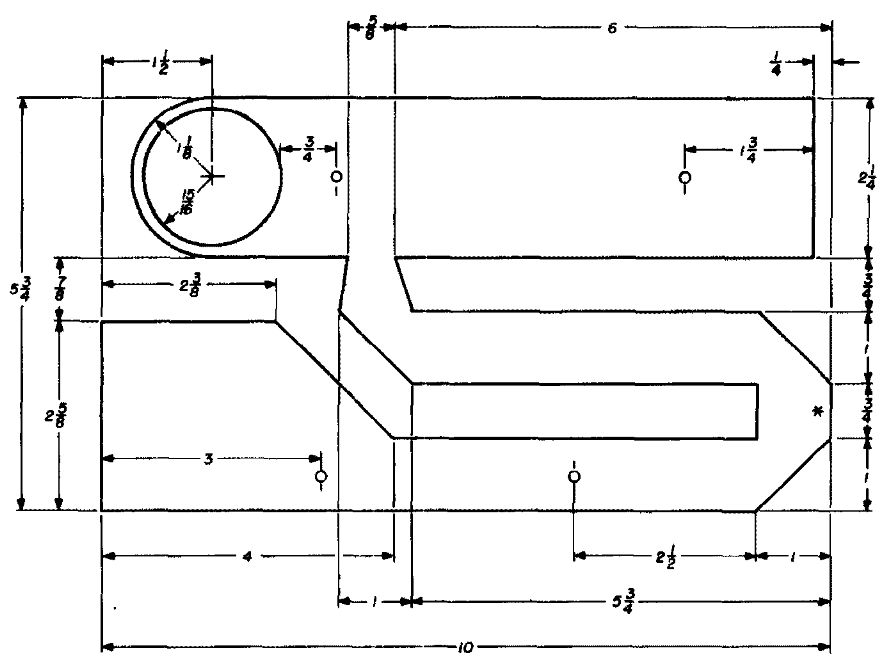
greater than 2 mils should work well. I don't recommend brass for the two-meter line or excessive use of solder to connect the rf choke at the high current point since these materials have higher resistivity.

The finger stock around the tube anode is soldered to both sides of the copperclad board. The intrinsic capacitance between the two sides of the board eliminates the need for any other connections. The rf choke is connected to the upper side only. I like the softer, folder-over type, silver-plated finger stock for ease of the tube removal, but the stiffer brass types will also work.

The stripline is supported above ground plane by four 1-1/2-inch long threaded porcelain spacers. Their location was chosen so that they don't lie in an area of high rf potential which might cause excessive dielectric heating, or in an area of high stripline current. They were also placed so that they hold the stripline rigid at the flapper capacitors, and no evidence of drifting has been noted during 144 or 432 operation. Rexolite or Teflon spacers might be superior electrically, but properly-placed porcelain ones seem quite adequate.

Cooling air is brought into the lower chassis through a 1-1/2-inch hole and out the upper chassis right side through a 1-1/2-inch screened hole. The air traverses the length of both chassis and cools all com-

**fig. 4. The dual-band stripline is built from 1/16"-thick double-sided 2-ounce copper-clad epoxy-fiiberglass printed circuit board. B+ is connected at the point marked with an asterik.**



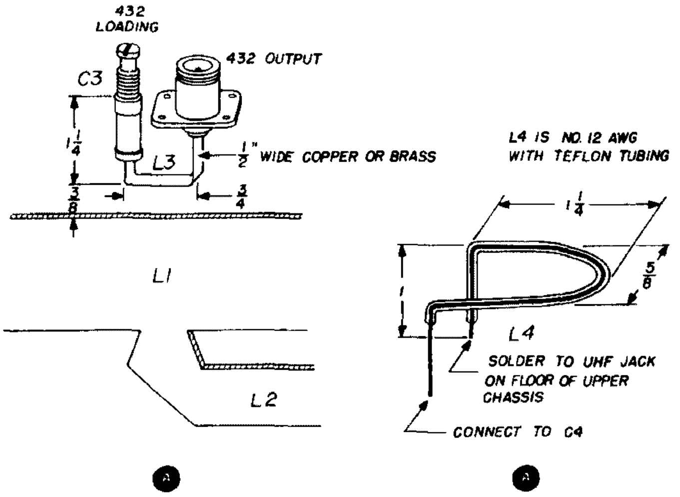



**fig. 5. Output coupling loops. Capacitor C4 is mounted on the front of the amplifier 2-1/8" below the top and 3-7/8" in from the right side. L4 loop is positioned ½" from inside bend of L2.**

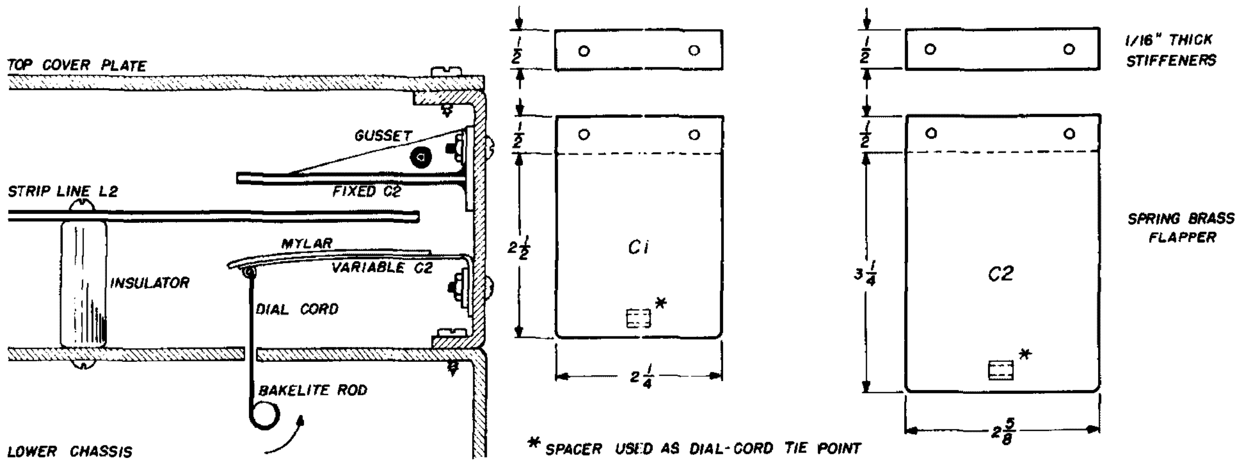
ponents. The screening was first tinned around the edges; the inside of the hole was tinned with aluminum solder using a soldering iron attachment on a butane torch. It was then easy to tack the screening in place with the iron attachment and torch.

## tuning capacitors

The flapper capacitors C1 and C2 are built as shown in **fig. 6.** They are made from spring brass; the dial cord is brought through a small hole in the chassis and wrapped around a bakelite rod which extends through the front panel. Drilling a small hole in the rod and tying the dial cord through it avoids any slippage. If the two holes through the chassis are small enough, there's almost no loss of cooling air. **Do not** use metallic dial cord.

The bakelite rods are each supported by two shaft bushings which are intentionally misaligned to cause friction so that the capacitor will hold its setting. A sheet of 10-mil Mylar film glued to the top surface of each flapper with silicone rubber adhesive insures that a high-voltage short will not occur if the flapper touches the stripline.

Capacitor C2 consists of the variable flapper on the bottom of the stripline and a fixed but semi-adjustable capacitor plate on the top side. It is properly adjusted when



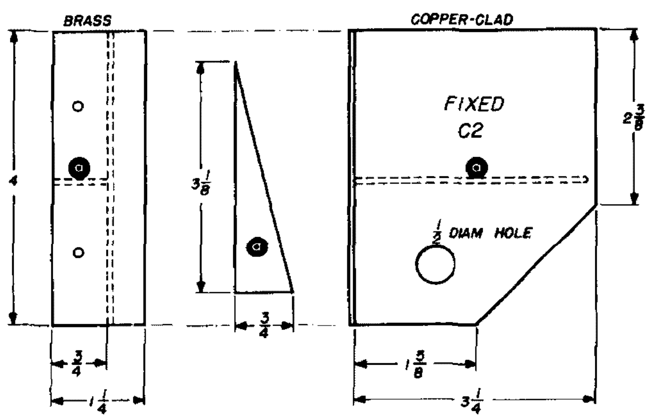




**fig. 6. Construction of variable capacitors C1 and C2. For installation, 6-32 nuts are soldered to the brass stiffeners and the mounting plate for the fixed portion of C2. The C1 flapper is located 3/8" below L1; C2 flapper is 1/4" below L2 and the fixed portion of C2 is 3/16" above L2. The fixed portion of C2 is shown bolted to the side of the chassis for illustration— it is actually bolted to te front of the chassis as shown in the photo. The 1/2" hole in the fixed portion of C2 is centered on the stripline mounting bolt.**

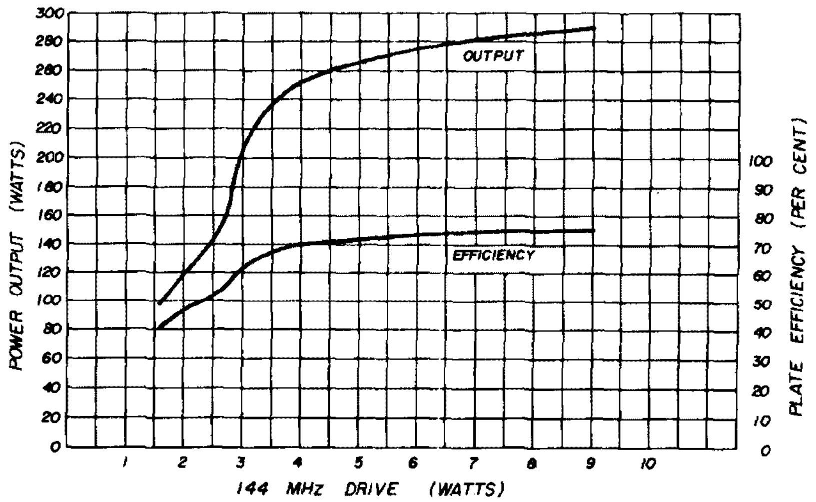
the two-meter line resonates with equal spacings of these two capacitor plates—approximately 3/16 inch.

## power supplies

A 500-watt input on 144 MHz requires 5.2 watts drive and +2000 volts at 250 mA, +270 volts at 30 mA and −90 volts bias; 80 watts output as a 432-MHz tripler requires than 375 watts output may be obtained. The −90V grid-bias supply makes adjustments easier and eliminates the need for B+ relays. When drive is removed, the tube turns off, even though B+ and screen voltage are still present.

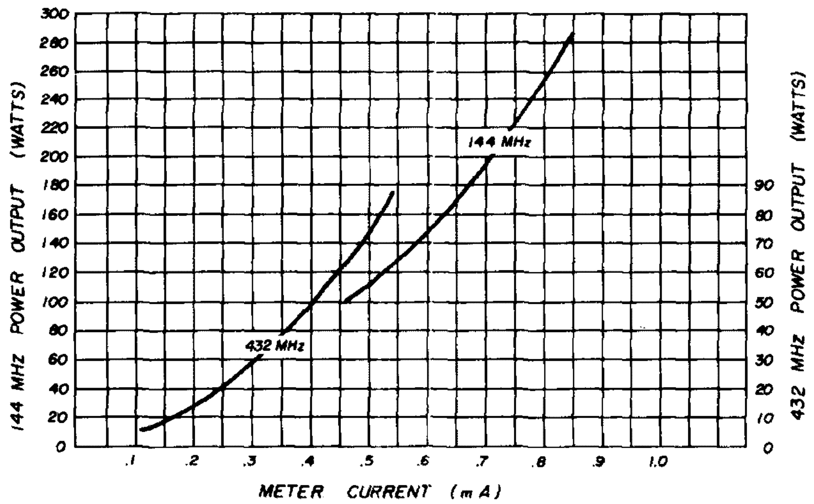
The 4CX250B filament is designed for 6.0 volts; a dropping resistor of 0.08 ohms must be used if a 6.3-volt transformer is

**fig. 7. Operating characteristics of the hot-carrier-diode output meter.**




20 watts of two-meter drive and 1400 volts B+. Less drive will still give acceptable results as **fig. 8** and **9** illustrate. These measurements were made with a used 4CX250B tube in the final. All power measurements were made with +1400 volts supply, the limit of my high-voltage power supply. If you use a 2000 volt supply for two-meter operation, efficiency will be slightly higher, and greater used. This 2-watt resistance is simply formed with a 5-foot length of no. 22 wire coiled around a suitable form. An optional wirewound potentiometer allows decreasing the 6.0-volt filament voltage to 5.5-V for 432-MHz operation. The back-bombardment of the cathode caused by 20 watts drive causes considerable heating, and decreased filament voltage will extend tube life.

**fig. 8. Operational characteristics of the amplifier/tripler when used as a 144-MHz amplifier with B+, 1400 V, screen voltage, 270 V, and grid bias, −90 V. Higher output and efficiency can be obtained with 2000 volts B+.**



On the subject of 4CX250B's, I have found that many of these tubes will blow the screen-grid and plate fuses when excitation is first applied. This may happen two or three consecutive times, and after that no trouble is experienced for hundreds of hours. I suspect that small pieces of material which seem to be floating around in some of the tubes occasionally lodge in a critical area. Each time a fuse blows, one of these pieces is vaporized.

I fuse the B+ with a 0.5 ampere 3AG fuse in a well insulated fuse holder connected in the negative side of the supply. The screen supply is fused with a 1/4-ampere 3AG fuse. The control-grid bias supply uses a current-limited regulator.

High voltage is brought into the chassis with RG-59/U cable and Amphenol MHV connectors. MHV connectors are rated to 5000 volts, and they automatically create a ground through the coax shield; RG-59/U cable is rated at 2300 volts rms. I strongly recommend using these or a similar type of high voltage connector for any final because groundless connectors represent a dangerous hazard if the ground system is disconnected.

## neutralization

The two-meter neutralizing circuit consists of a 3/4-inch diameter brass plate held 1/2-inch below L2 in the vicinity of C2 by a stiff piece of wire soldered to a Teflon feedthrough. A similar stiff piece of wire on the other side of the chassis connects the feedthrough to piston capacitor C8. The rotor of C8 is fastened to a large ground lug which is soldered to the grid pin of the 4CX250B.

The folding of the two meter cavity reduces the total length of the neutralizing circuit. This prevents the need for floating the grid circuit components, a technique which is required with the usual two meter neutralizing circuit. It is reassuring to perform the neutralizing procedure and observe the power fed through to the plate circuit disappear by approximately 30 dB.

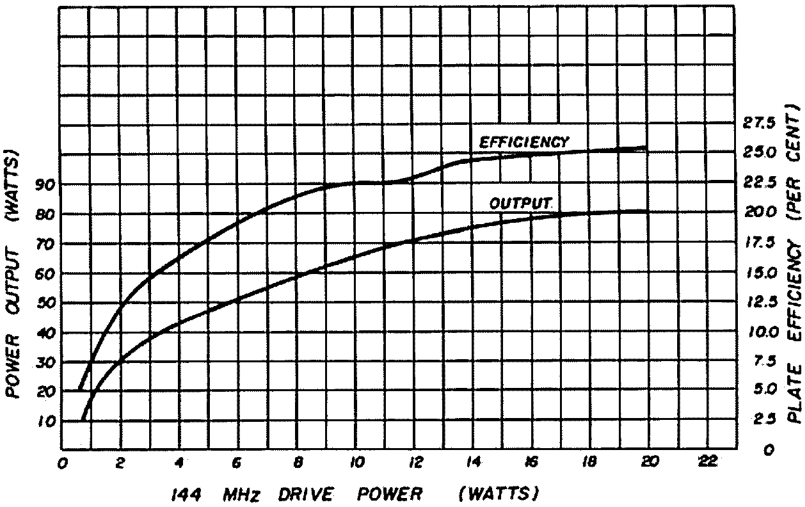
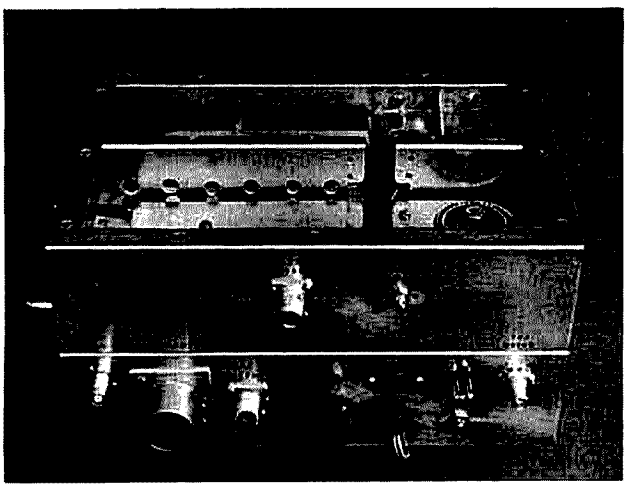



**fig. 9. Operational characteristics of the amplifier when used as a tripler from 144 MHz to 432; B+, 1400 V; screen voltage, 270 V; grid bias —90 V.**



**Rear view of unit shows high-voltage B+ MHV connector on lower left. The near cavity with the 4CX250B is tuned to 432 MHz; far cavity is tuned to 144 MHz.**

Neutralization is carried out with filament and grid-bias voltage turned on and an appropriate amount of two-meter grid drive applied through an swr meter to J1. As you adjust C5 and C6 to 1:1 swr you will note a peak in grid current if your driver has a 50-ohm source impedance. If no swr meter is available merely adjust C5 and C6 for maximum final grid current.

Next, minimize C1. A grid-dip meter in the diode position and tuned to 144 MHz is held near the bend in the two-meter line, and C2 is adjusted for maximum indication. An insulated tuning tool is used to adjust C8 for minimum indication on the grid-dip meter. The amplifier is now neutralized. Slightly more accurate neutralization is obtained if the top cover is screwed dowi and the grid meter is coupled to a two turn link at J3. Very stable two-meter operation occurs with either neutralizing scheme.

## tune up

The two-meter output circuit is adjusted with S1 and S2 in the 144 position and adjusting C2 and C4 for maximum indication on the output meter. L4 is adjusted for proper loading by bending it away from or nearer L2. When proper loading is obtained, C4 will be near its mid position, and screen current will be approximately 15 mA. Screen current should be monitored closely during tuneup since it's the best indication of tube loading: increasing screen current indicates decreased plate loading.

Screen current in excess of 40 mA for any appreciable time may destroy the screen grid. The plate, by comparison, is quite rugged when air is forced through it at greater than 10 cubic feet per minute. Air pressure at 0.5 inch of water is required to do this. A 4-inch squirrel-cage blower should be used.

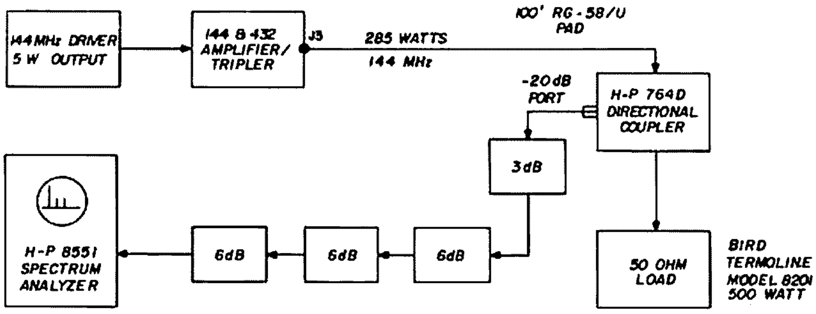
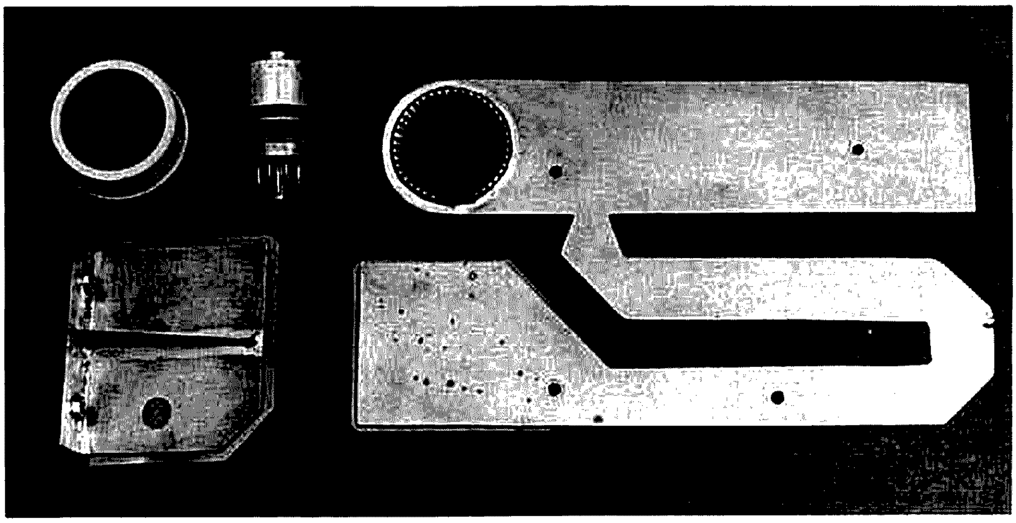



**fig. 10. Test setup for the spectrum display of the 144-MHz ouput. Spectrum display is shown to tho left. Center of display is 500 MHz, horizontal scale is 100 MHz/cm and vertical scale is 10 dB/cm.**



**Resonant components are held in place with six screws. Stripline is to the right and fixed portion of C2 is on lower left; 4CX250B and air chimney are in upper left.**

During 432-MHz tuneup 144-MHz grid drive is increased to 20 watts—if you have that much—switches S1 and S2 are set in the 432 position, C2 is minimized and a suitable load is connected to J2. When maximizing 432 output there is usually no clearly defined plate current dip; this is a normal characteristic of triplers, but the output metering circuit makes adjustment very easy. C1, C3, C5, C6 and R2 are varied while observing the output meter. I have found that lowering the screen or plate voltage is the best way to decrease output on 432. Decreasing the loading merely decreases the efficiency.

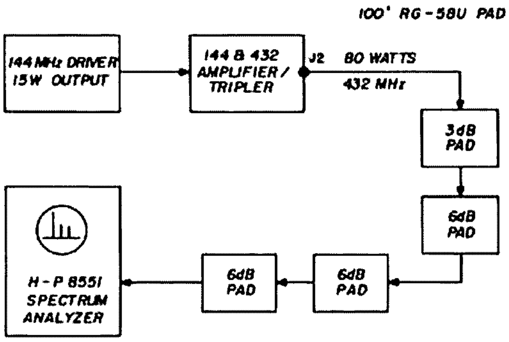
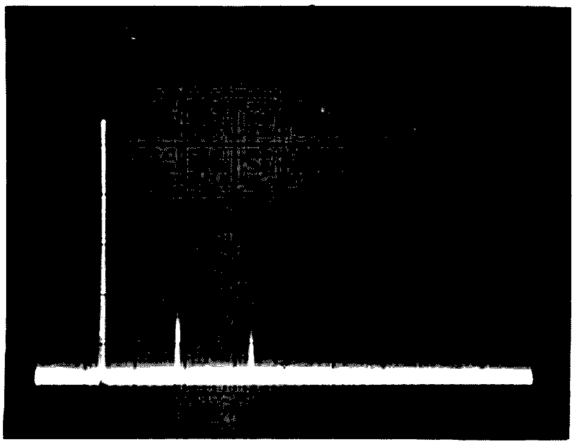
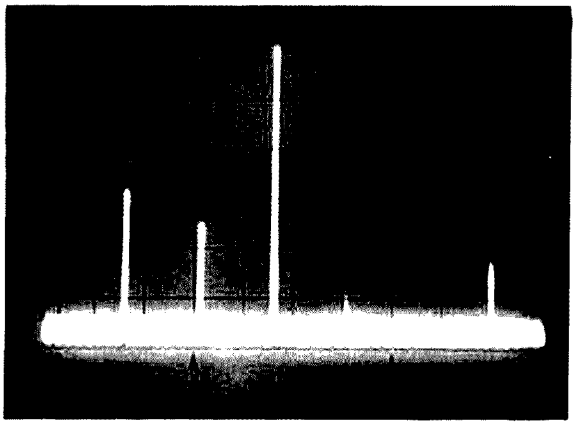



**fig. 11. Test setup for the spectrum display of the unit operating as a tripler from 144 to 432 MHz. Spectrum display is to the right. Center of display is 480 MHz, horizontal scale is 100 MHz/cm and vertical scale is 10 dB/cm.**



**Spectrum analyzer display of 144-MHz output.**



**Spectrum analyzer display of 432-MHz output.**

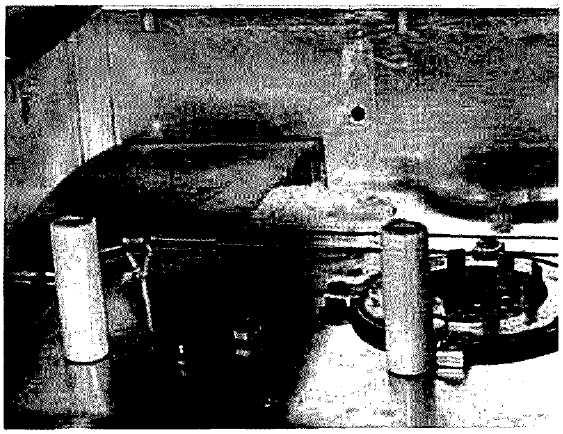
Schottky-barrier diodes with a recovery time of 100 picoseconds make a simple 432-MHz output metering circuit possible without a microammeter. Maximum rf voltage output always yields maximum power output regardless of the standing wave ratio of the antenna or the length of the transmission line (if no changes in the antenna or transmission line occur during the tuneup). The output-meter characteristics are shown in **fig. 7.**

## performance

Complete performance characteristics of the 144/432 MHz amplifier/tripler are shown in **figs. 8** and **9.** On 144 MHz, power output is 285 watts with 9 watts drive and a 1400-volt plate supply. Efficiency is a respectable 75 percent. Higher output could be obtained with a higher B+ supply. As a tripler to 432 MHz, the unit provides 80 watts output with 20 watts of two-meter drive; efficiency in this mode is about 25 percent.

One of the most important characteristics of vhf power amplifiers and frequency multipliers, and one of the least discussed, is the level of spurious outputs. Spectrum displays of this amplifier/tripler's output on 144 and 432 MHz are shown in **figs. 10** and **11** along with the test setup used.

When operated as a power amplifier on 144 MHz, 288 MHz output was 41.5 dB down, 432 MHz output was 40.5 dB down and 576 MHz output was more than 50 dB down. With 285 watts output on two



**Closeup shot shows construction of C2.**

meters, this represents 20 mW on 288 MHz, 25 mW on 432 MHz and less than 3 mW on 576 MHz.

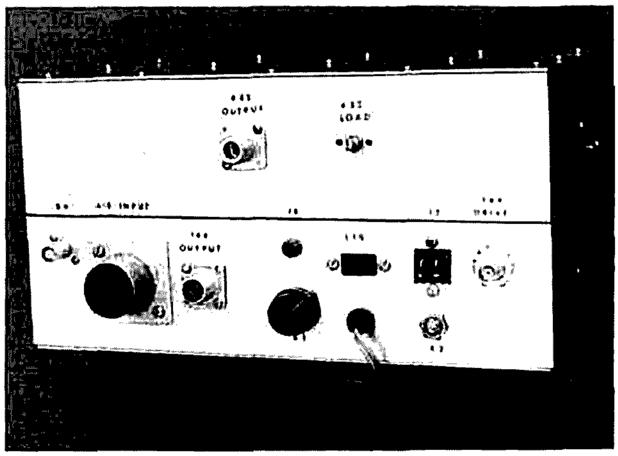
As a tripler from 144 to 432 MHz, 144 MHz output is 34.0 dB down, 288 MHz output is 38.3 dB down, 576 MHz output is 50 dB down, 720 MHz output is 65 dB down and 864 MHz output is 43 dB down. With 70 watts output on 432 MHz, these figures correspond to 27.8 mW on 144, 10.4 mW on 288 MHz, 0.7 mW on 576 MHz, 0.02 mW on 720 MHz and 3.5 mW on 864 MHz. These figures have all been corrected for cable loss; in addition, the 144-MHz amplifier figures include measured directional coupler losses on each of the frequencies of interest.

## summary

The techniques used in this dual-band final have other applications. With modifications of the grid and plate circuits, multiband operation on 144, 220 and 432 MHz should be possible. If a dual-band grid circuit similar to the present plate circuit is constructed, then this final will also have the option of operating straight through as an amplifier of 432 MHz. A multiband grid circuit doesn't need to be nearly as large as the plate circuit since much smaller power handling is required. The flapper capacitors could be replaced with commercial variable capacitors.

Most of the modern rf developments seem to be occurring in the solid-state field, but a person familiar with both vacuum tubes and transistors can select the best techniques of each discipline. Although stripline was primarily a transistor-oriented development its advantages should give it considerable application with vacuum tubes. Stripline design graphs are available in "Reference Data for Radio Engineers."[6]

This amplifier/tripler would not have been possible without consultations with stripline authority Henry Keen, W2CTK.[7] Thanks also go to James Buscemi, K2OVS, Edward Mentz, K2LCK, Rchard Winderman, WA2OBG, Fred Telewski, WA2FSQ and Douglas McGarrett, WA2SAY for their valuable assistance.



**Rear view of the complete amplifier/ tripler unit for 144 and 432 MHz.**

### references

1. William I. Orr, W6SAI, John T. Chambers, W6NLZ, "Stripline Kilowatt for 432 MHz," *ham radio*, September, 1968, p. 6.
2. Robert I. Sutherland, W6UOV, "Design Data for A Two-Kilowatt VHF Linear," *ham radio*, March, 1969, p. 6.
3. Frank C. Jones, W6AJF, "VHF for the Radio Amateur," Cowan Publishing Corporation, 1961, pp. 81, 82, 90, 97.
4. S. Cohn, "Characteristic Impedance of the Shielded-Strip Transmission Line," *IRE Transactions*, Vol. MTT-2, July, 1954, pp. 52-57.
5. Frank Davis, "Matching Network Designs with Computer Solutions," Motorola Application Note AN-267, 1967, Motorola Semiconductor Corporation, Phoenix, Arizona.
6. H. P. Westman, editor, *Reference Data for Radio Engineers*, 1956, ITT Corporation, New York, p. 598.
7. H. S. Keen, "Scientific Report on Study of Strip Transmission Lines," *Airborne Instruments Laboratory Report* No. 2830-2, December 1, 1955.

**ham radio**

# phase-modulated transmitter for two meters

An interesting design featuring solid-state devices, narrow bandwidth, and low power comsumption

Frank C. Jones, W6AJF, 850 Donner Avenue, Sonoma, California 95476

**The transmitter shown here** can be operated as a complete low-power vhf unit, or it can be used to drive a larger amplifier. The output at 144 MHz is between one-half and one watt when used with a 12-volt battery. As much as two watts can be obtained by running the buffer and final amplifier from a 26-volt supply, but this is getting toward the upper limit of the type 40280 transistor. The 40280 has a maximum rating of 7 watts input at 25°C, so under normal conditions, even with a good heat sink, about 4 watts input or less would be a safe limit.

## transistor ratings

The buffer and last two doubler stages use a new, low-priced (55 cents) RCA type 2N5188 transistor rated at 4 watts maximum at 25°C. Probably two watts input would be a safe rating with a moderately good heat sink, and one watt input with only a clip-on collector and case cooler or radiator. Two different types of heat sink are used, as shown in the top chassis view. The black large-finned type is a little better for cooling than the small-ribbed aluminum coolers. The RCA 40280 output amplifier was slipped into a stud-mounting casing and the casing was mounted in a large-finned heat sink with mica washers for insulators. This permits operation at higher voltage and current, but it is not essential, as a 12-volt supply is used for all stages. A fairly good-sized finned cooler would be sufficient in this case if the input to the amplifier is kept to 1½ watts.

The 40280 transistor costs nearly five dollars as compared to fifty-five cents for the 2N5188, but is about 20 to 25 percent more efficient at 144 MHz and has a greater input power rating. If a half-watt output is sufficient, a 2N5188 can be used in the output stage at considerable savings in cost. Some 2N5188 transistors are apparently

nearly as good as the 40280. Probably their $f_T$ is higher than average and approaches that of the 40280. In trying out a dozen 2N5288's, two were found that worked extremely well at 144 MHz, and three others were well above average in output. All functioned very well up to 72 MHz and not too badly at 144 MHz with a 12-volt supply. They have a maximum collector-to-emitter rating of 25 volts and a base-to-emitter rating of 5 volts. These values should never be exceeded, and a safe figure of 1/2 of these values is advisable for dc ratings.

The 40280 has similar maximum ratings of 36 and 4 volts, so both types should operate from a 12- or 13-volt battery with long life expectancy. The 40280 has an $f_T$ of 550 MHz vs 325 MHz for the 2N5188, which means less efficiency at vhf for the lower-priced transistor. Both are double diffused, epitaxial planer silicon transistors, with the same case and lead arrangements. The input and output capacitances are a little different, which means retuning circuits when interchanging these two types.

The 2N3553 is similar to the 40280 in cost and performance, but can be used at higher supply voltages, since its maximum rating is 65 volts. At 12 volts the results were about equal to the best 2N5188's.

## amplifier stability

The amplifier circuit used with the 40280 was regenerative, so that care had to be taken to tune the input and output circuits correctly for maximum output at only the desired frequency. Using a dummy antenna, the rf output would sometimes double with the same rf drive, but with more than double the collector current. Monitoring the output with a receiver indicated faulty operation with lots of "hash" noise and less-than-normal carrier output at the desired frequency. This condition seemed to take place when the series output tuning capacitor was set at high values of capacitance. When kept below 15 pF, normal operation occurred with 100 to 150 mA at 12 volts.

## the importance of high Q

Unless extremely high-Q tuned circuits are used, a single interstage circuit will have more loss and less harmonic suppression than two circuits of moderate unloaded Q. These stages can then be loaded down to an operating Q of 10 or so, with good efficiency and good harmonic suppression. This approach was used in the transmitter illustrated here. Double-tuned circuits were used in all doubler and vhf stages.

## frequency multiplier efficiency

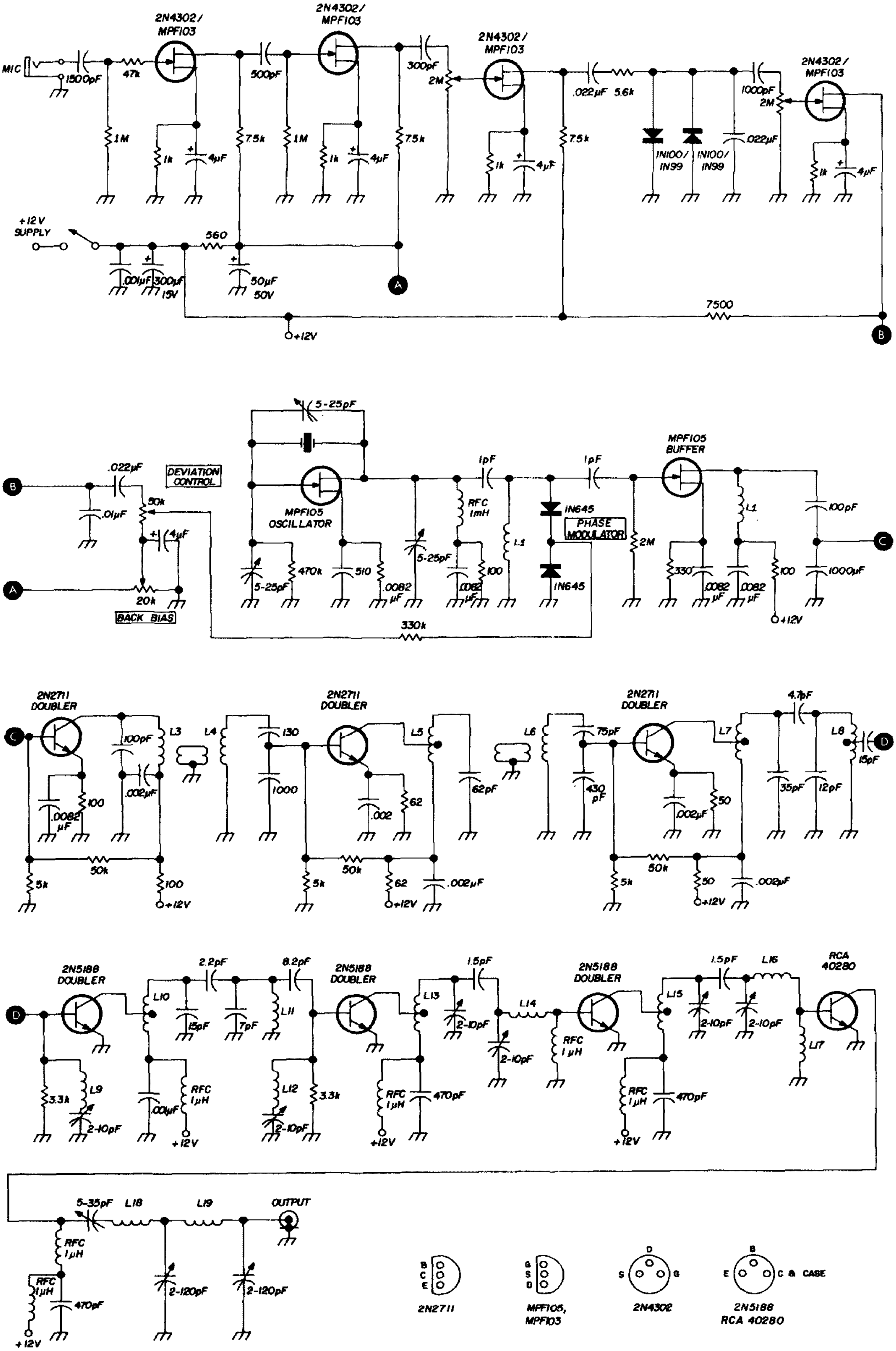
Two types of frequency doublers were used, both having a low base-to-emitter impedance path for the doubler output frequency. Transistor doublers have considerable output-to-input feedback through the base-to-collector capacitance. This required a low impedance path from base to emitter to get much efficiency in the frequency multiplier.

The methods used here consisted of a low C-to-L ratio, series-tuned circuits from base to ground or emitter for the vhf range. A capacitance divider for matching the low base-input impedance to a higher LC impedance also provides a fairly low base-to-emitter impedance at the harmonic frequency as compared to the capacitance divider circuit.

Measurements indicated about twice as much output power available for the same rf drive when the base-to-emitter impedance is very low at the output harmonic frequency. No tests were made here in that regard for straight-through rf amplifiers. The capacitance divider coupling circuits for base input were used in the lower rf ranges.

Low-priced 2N2711 or 2N5182 bipolar transistors were used in the first three doubler stages. These have different pin connections, but are not too unalike for doubler service in this transmitter. Older type 2N706's can also be used, but these have a lower beta and are not quite as efficient as the newer types. In general, the higher-numbered transistors are more efficient and cost less (2N5182 is 40 cents new).

The circuit diagram (**fig. 1**) and photograph show plastic 2N2711's, since these were available from another project of a few years ago. The 2N706's were also tried in the last doubler and in the 144-MHz buf-



MIC
1500pF
47k
2N4302/
MPF103
1M
1k
4μF
500pF
7.5k
300pF
2M
.022μF
5.6k
1N100/
1N99
1000pF
.022μF
+12V
SUPPLY
560
.001μF
300μF
15V
50μF
50V
A
7500
+12V
B
5-25pF
DEVIATION
CONTROL
.022μF
.01μF
50k
20k
BACK BIAS
MPF105
OSCILLATOR
470k
510
.0082
μF
1pF
RFC
1mH
100
L1
1N645
PHASE
MODULATOR
2M
MPF105
BUFFER
330
100pF
1000μF
330k
C
2N2711
DOUBLER
L3
L4
130
1000
.002μF
62
.002
L5
L6
62pF
75pF
430
pF
50
L7
4.7pF
L8
35pF
12pF
15pF
5k
50k
D
2N5188
DOUBLER
L10
2.2pF
15pF
7pF
8.2pF
L11
L9
3.3k
2-10pF
RFC
1μH
L12
L13
1.5pF
470pF
L14
L15
L16
RCA
40280
L17
5-35pF
L18
L19
OUTPUT
2-120pF
2N2711
MPF105,
MPF103
2N4302
2N5188
RCA 40280
C & CASE

**fig. 1. Schematic of the 144-MHz solid-state phase-modulated transmitter is shown to the left. Low-cost transistors and high-Q circuits are featured.**

| | |
|---|---|
| **L1** | **Broadcast band loopstick, 100 to 250 μH.** |
| **L2** | **40 turns no. 30E, ½ inch long x ½ inch diameter. $F_r$ slug.** |
| **L3, L4** | **22 turns no. 28E, 5/16 inch long x ½ inch diameter. $F_r$ slug. 3½ turn links.** |
| **L5** | **17 turns no. 26E, 5/16 inch long x ¼ inch diameter. Tap 10 turns up. 2½-turn link.** |
| **L6** | **Same as L5, except no tap. 2½-turn link.** |
| **L7** | **12 turns no. 26E, ¼ inch long x ¼ inch diameter. Brass slug. 6-turn tap.** |
| **L8** | **Same as L7, except no tap.** |
| **L9** | **21 turns no. 24C, ½ inch long x ¼ inch diameter. No slug.** |
| **L10** | **9 turns no. 22E, ¼ inch long x ¼ inch diameter. Brass slug. Tap 3 turns up.** |
| **L11** | **Same as L10, except no tap.** |
| **L12** | **10 turns no. 24C, ¼ inch long x ¼ inch diameter. No slug.** |
| **L13** | **5 turns no. 18E, ½ inch long x 5/16 inch diameter. Tap 2 turns up.** |
| **L14** | **8 turns no. 18E, 3/8 inch long x ¼ inch diameter.** |
| **L15** | **5 turns no. 18E, ½ inch long x 5/16 inch diameter. Tap 1½ turns up.** |
| **L16** | **5 turns no. 18E, ½ inch long x ¼ inch diameter.** |
| **L17** | **3 turns no. 18E, 3/16 inch long x 3/16 inch diameter.** |
| **L18** | **6 turns 1/16-inch wide copper strip or no. 14E, 1 inch long x 5/16 inch diameter.** |
| **L19** | **Same as L18, except 4 turns.** |

fer stages, but they got pretty hot and were less efficient than the 2N5188's. These stages were biased and driven to about 50 mA of collector current with a 12-volt supply. A dc voltmeter with an rf choke in one lead should be used to check the voltage on the doubler stages. The bias resistor in the base circuit should be low enough to keep operation well below the maximum rated base-to-emitter values listed in the transistor data sheets or handbooks. Too low a value will reduce rf output.

## phase modulator

The phase modulator is rather interesting. It is a high-Q, very low-C circuit consisting of a broadcast band "loopstick" slug-tuned coil with an inductance range of about 100 to 250 μF. It is coupled lightly to the crystal oscillator and to the buffer input through 1-pF capacitors to keep the circuit at a high Q and low C.

Two small silicon (200 mA, 200 V) type 1N645 diodes were connected back-to-back, with some reverse dc bias, to act as small variable capacitors across the loopstick coil. The resting dc bias, of some value between 6 and 12 volts, sets the initial capacitance of these "variable capacitor diodes." The voltage from the audio amplifier then varies the capacitance and the phase of this circuit.

The coil slug is adjusted to bring the circuit near the crystal frequency, since the LC phase changes rapidly and fairly linearly in this region. The variable-capacitor diodes, back-biased, have a resistance of probably 1000 megohms; thus they have practically no effect on the circuit Q. A reactance-type transistor would load the circuit. The diodes are connected back-to-back to reduce the possibility of the peak rf voltage biasing the diodes in the forward direction.

Very small capacitance coupling to the oscillator is needed to keep the rf peak voltage to a few volts across this high-Q circuit. The higher the Q, the less capacitance change needed for a given amount of phase modulation.

## modulator diode considerations

A low-Q circuit in the modulator usually requires special varicap diodes, which are more costly than small silicon power-rectifier diodes. Computer diodes generally have too small a capacitance change to work well in this circuit. "Top hat" rectifier diodes also work, but often have a larger shunt capacitance, so the small glass-enclosed diodes are better. Pick a couple that show "infinite" back resistance on an ohmmeter test, and be certain that the cathodes are connected together and to the af and plus dc bias circuit. The power rectifier diodes have much less capacity change than a regular varicap, but work in exactly the same manner.

The af amplifier uses some surplus fet's in a resistance-coupled system. Motorola MPF103's are reasonably priced and may be used in place of the 2N4302 transistors shown in **fig. 1.** The socket connections are different, however.

A simple diode speech clipper was used to run the phase modulation at a fairly high average level. This requires an additional af stage and gain control to function well.

## construction notes

At this point it should be apparent that this article is more concerned with ideas for good transistor circuit design than with a pure mechanical layout or design that can be copied exactly. As the photographs show, there are extra holes in the 5x13-inch copper-plated panel that resulted from numerous experiments in the design. This panel, copper side down, was mounted in an old 5x13x3-inch chassis for shielding and protection of parts. Coil data is given in **fig. 1.**

After some difficulties were ironed out in the crystal oscillator, this transmitter was used to transmit fm signals through a MARS repeater station. The main problem here was to get onto the exact frequency and stay there. One of the npo variable ceramic capacitors in the crystal oscillator either was not really an npo unit or had a faulty bearing. Once the correct values were obtained, the ceramic variables were replaced with fixed silver mica capacitors except for the one across the crystal (this was changed to an air-type variable padder so the crystal could be set to oscillate within a few Hz of the value desired). A 22 pF capacitor from gate to ground and a 5 pF or 10 pF from drain to ground were suitable values. A 5- to 25-pF variable was placed across the crystal. Motorola MPF105 fet's were used in the crystal oscillator and in the buffer stage at 4.6± MHz.

## operation and tuning

The five doubler stages multiply the crystal frequency 32 times and increase the phase modulation. The final result is a phase-modulated signal in the two-meter band with a bandwidth of 5 kHz or more, suitable for voice communication with a narrow-band fm receiver.

The total current drain with a 12-volt dry battery, storage battery, or regulated power supply is about 1/3 ampere. If the current is near 0.5 A, probably a parasitic oscillation is present in the 40280 transistor stage.

In tuning up the transmitter, the total current drain will be about 10 or 20 mA with the crystal out. Each stage, as it is tuned to resonance (with the crystal oscillator working), will increase the total dc current a few mA until the power transistor stages are reached. All these stages, using bipolar transistors, run at near-zero current until rf drive is applied to each stage. If any rf or doubler stages in this system draw appreciable current with the crystal out, then parasitic oscillation, incorrect wiring, or faulty transistors can be suspected.

After some rf output is obtained, the series-trap circuits at 72 and 144 MHz can be peaked for maximum output. Some tuning adjustments on these circuits can actually reduce the stage output to a very low value. The correct adjustment results in maximum rf output.

A radio receiver is used to monitor the output on two meters during tuneup. Use a 50-ohm, 1-watt dummy antenna and rf voltmeter (or a dummy antenna and swr meter). Again monitor the output on two meters after an antenna is connected, since an antenna usually presents a different load than a dummy antenna. This may cause the output amplifier to self-oscillate until proper loading is again set in the output pi coupling circuits. Physical layout, ground connections, bypass capacitor reactances and rf chokes can all cause variations in the operation of an unneutralized rf amplifier.

Phase or frequency modulation permits more leeway than if the transmitter were amplitude modulated. Personally, it seems to me that a transistor linear amplifier would be easier to tame than an amplitude-modulated amplifier. Even in a cw or phase-modulated transmitter, parasitic or self oscillation is definitely not to be permitted on the air.

**ham radio**

# a survey of solid-state power supplies

Regulated power is a must for today's new circuits— here's a report on some new devices and their applications

Hank Olson, W6GXN, P.O. Box 339, Menlo Park, California 94025

**Since the publication of a** previous article on power-supply design[1], the semiconductor industry has introduced devices that have made a tremendous impact on the world of power supplies. Therefore, a review of power supply theory, with a transition to these new semiconductor devices, is in order.

## basic circuit

The rectifier-filter section of the power supply is ordinarily one of the five familiar types shown in **fig. 1.** Of these, I don't generally use any but the full wave, bridge or conventional doubler. I favor these because they have much better regulation and ripple reduction. Ripple reduction is better because the ripple frequency is twice that of the line frequency. Compared to a half-wave circuit, the chokes and capacitors in a full-wave rectifier effectively have twice the inductive reactance and one-half the capacitive reactance respectively.

Another convincing argument in favor of the full-wave and bridge circuits is that either choke input or capacitive input may be used. Thus, for any given transformer, two dc output voltages are available in the bridge connection, and two more are available in the full-wave circuit.

I use the half-wave types (half-wave and cascade doubler, **fig.** 1) only when it's necessary to ground one side of the transformer secondary. Such a requirement usually occurs when adding on to existing equipment.

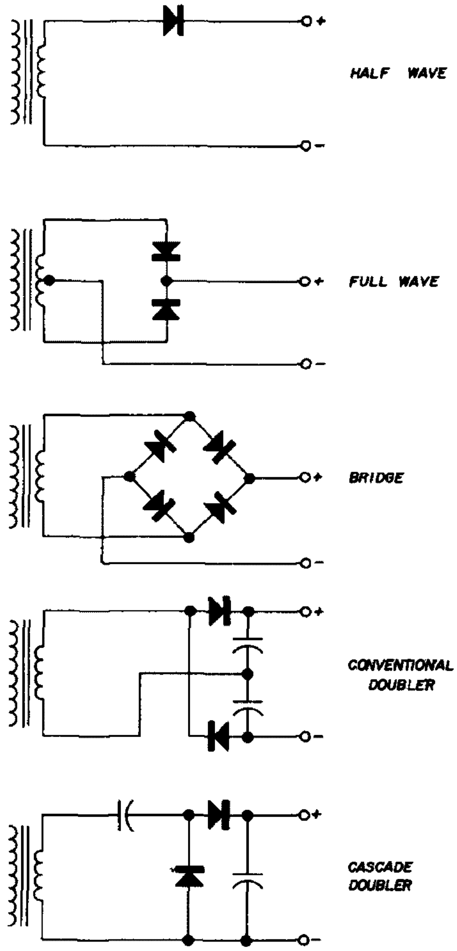



**fig. 1. Commonly used rectifier circuits. Full-wave versions provide best regulation and ripple attenuation.**

## series voltage regulators

The simplest voltage regulator that uses gain is the emitter-follower type, **fig. 2.** The transistor current gain, $h_{fe}$, allows this series regulator to regulate much more current than the zener by itself. Another important advantage is that the circuit also functions as a capacitance multiplier. The base voltage is filtered by R1-C1 in addition to any filtering ahead of the regulator. Since it is an emitter-follower, the emitter "follows" the base, so the output is as well filtered as the base.

Where did the ripple go? It's soaked up by the collector-to-base potential, which varies at the ripple-frequency rate. Note that enough voltage must be across the collector-base junction so that negative swings of the ripple voltage don't go below the zener-regulated base voltage.

## transistor current gain

The transistor current gain is important in this type of regulator as well as in the others to be discussed. The higher the current gain, the less base current necessary for any given regulated output current. The higher $h_{fe}$ is, the larger R1 can be. This results in better RC filtering (by capacitance multiplication). **Fig. 3** illustrates this with three similar emitter-follower regulators, their respective peak-to-peak ripple output, and output voltage as a function of current. All have the same rectifier-filter sections and load resistances.

## the darlington pair

**Fig. 3C** uses a trick called the Darlington pair: two transistors are connected as one to obtain the product of their current gains as the combined $h_{fe}$. This may seem unfair while we're making comparisons, but **this** Darlington pair is in one package and is used just like any other transistor. Of course, separate transistors can also be used similarly but the common-chip types have a thermal-tracking advantage. Note that there are two emitter-base diode voltage drops between the zener and the output, so a higher-voltage zener must be used.

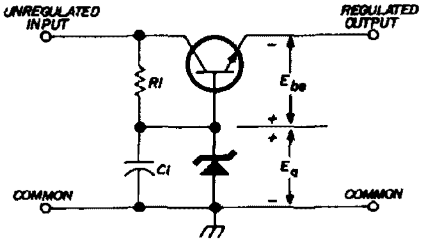



**fig. 2. Emitter-follower series regulator. Transistor current gain gives more current regulation than the zener alone.**

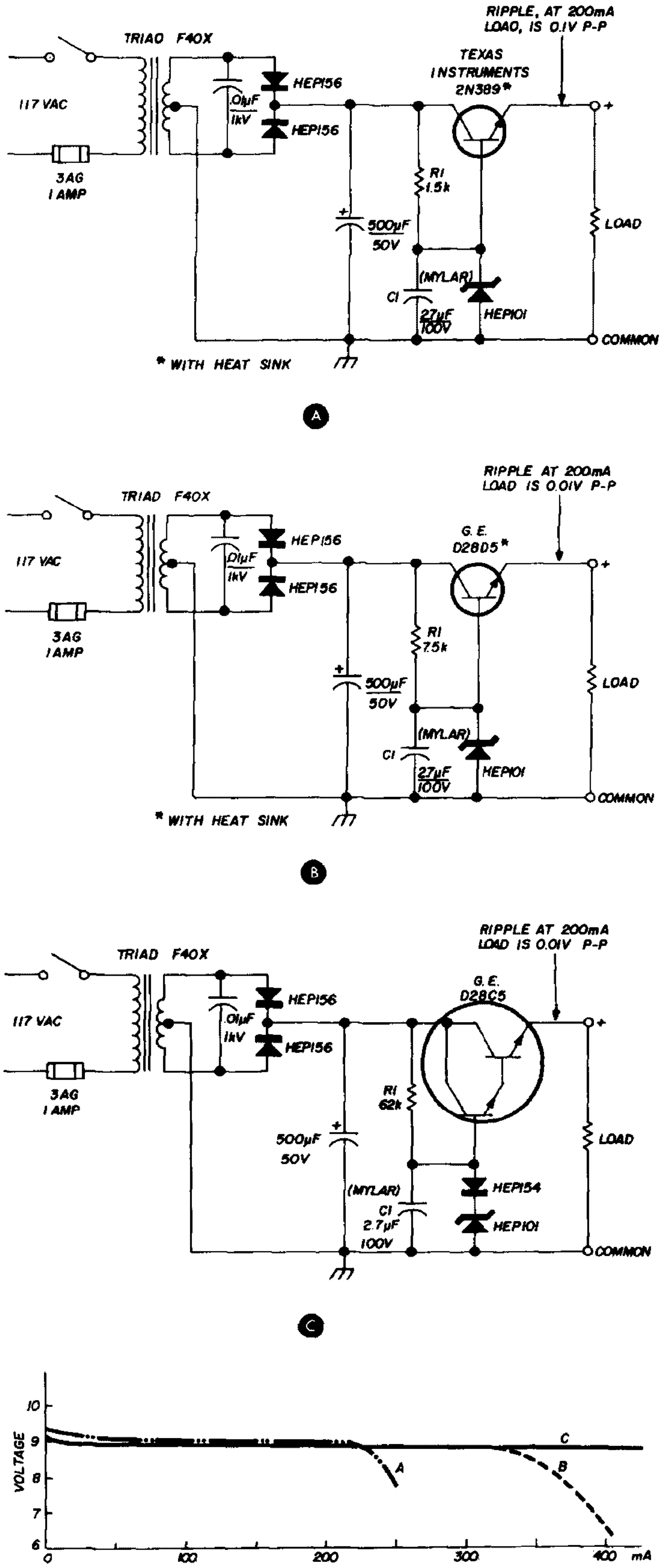



**fig. 3. Series regulators with "capacitance multiplication." The Darlington Pair is shown in C; higher zener impedance causes ripple voltage to be the same as in B.**

It's also possible to use another form of two-transistor combination in the emitter-follower regulator. **Fig. 5** shows such an arrangement; note that complementary transistors are used. Unfortunately, no single-package npn-pnp transistor is yet available for this circuit, so separate transistors were used. The advantage of this "compound emitter follower" over the Darlington Pair is that it has only **one** base-emitter diode voltage drop between control base and output.

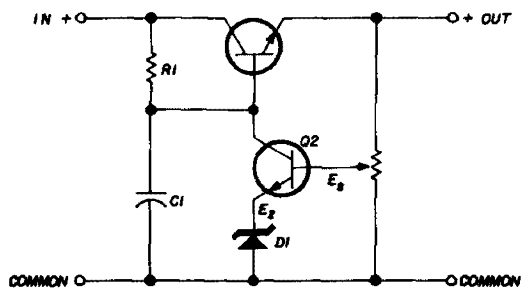



**fig. 4. Reference amplifier regulator, allowing adjustable output voltage. R1 is collector load resistance for Q2, a dc amplifier.**

## adjustable output voltage

The emitter-follower regulator, with its simplicity, is a useful regulating circuit, but it has a troublesome characteristic. The output voltage is fixed at $E_z - E_{be}$ (the zener breakdown voltage less the transistor base-to-emitter diode forward voltage drop.

By making the circuit a bit more complex, output voltage can be made adjustable. **Fig. 4** shows the more sophisticated circuit. Now, R becomes a collector load resistance for $Q_2$, a dc amplifier. $Q_2$ amplifies the difference between $E_z$, the zener voltage and $E_s$, a fixed fraction of the regulated output voltage (error voltage). This form of voltage regulator is widely used because it is simple and adjustable. Because of this wide use, several companies have offered integrated packages containing Q2 and D1[2,3]. An actual circuit is shown in **fig. 6A** for a "handle on reality." Note that in such circuits C1 is not made large for capacitance multiplication, but the added gain of the circuit is relied upon for ripple attenuation.

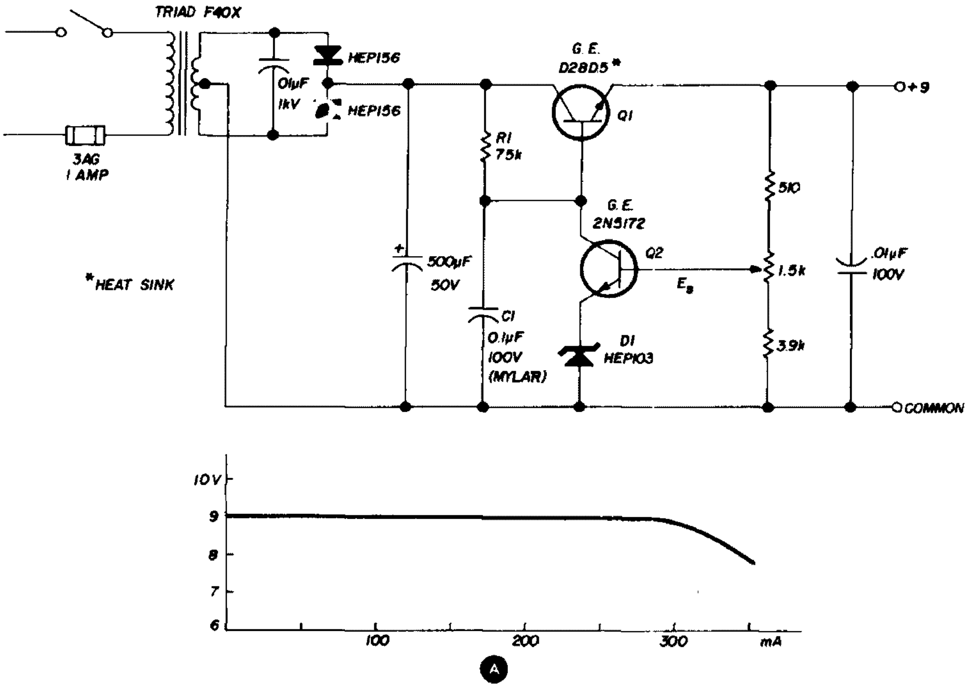



**fig. 6. Practical circuits using reference amplifier principle. The new G.E. "programable zener," a small-scale**

## the programable zener regulator

An interesting new device from General Electric is a monolithically constructed regulator called a "programable zener." It doesn't have a number assigned as yet, but if it's like other members of G.E.'s small-scale integrated circuit line, it'll probably have a D13XX number with a price tag of about a dollar. A circuit using it, **fig. 6B,** is similar to that of **fig. 6A.** Note its new circuit symbol. (I assume that the zener in this IC is an emitter-base breakdown diode.) The circuit using the programable zener regulator shows good regulation to well over one-half ampere of load current.

## differential dc amplifier regulator

By adding one or more transistors to our regulator, even better regulation can be obtained. **Fig. 7** shows how a differential dc amplifier allows the zener reference to run at a constant low current, improving reference stability.

The improvement offered by this differential amplifier is especially noticeable if its main function is to compensate for variations in the input voltage. A typical case would be where the input is from a 12-volt line in an auto electrical system. **Fig. 8** shows a practical regulated supply using the differential amplifier. The differential amplifier IC is one of the most practical monolithic circuits.

## fet regulator

Still another interesting development on the power supply scene is the use of power

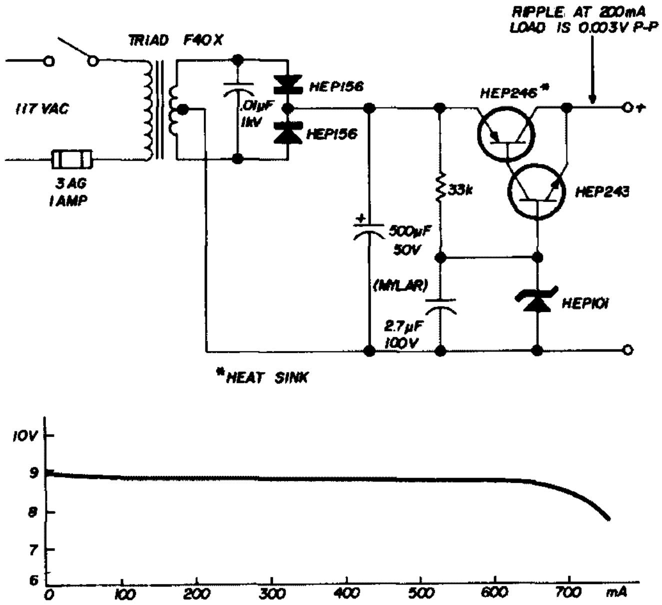
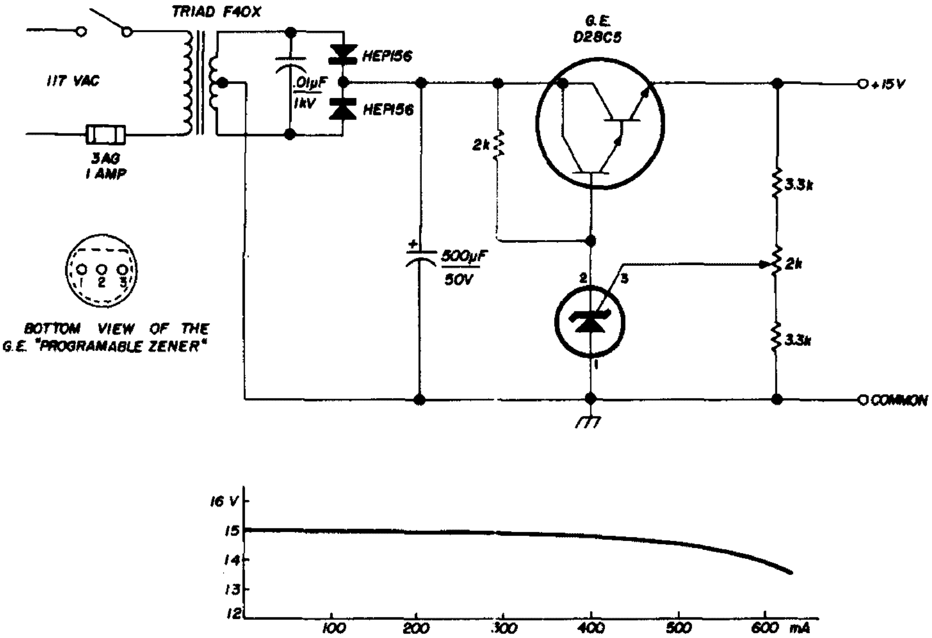



**fig. 5. The "compound emitter-follower" regulator. Compared to the Darlington Pair, it has only one base-emitter voltage diode drop between base and output.**




8

IC, is shown in B.

fet's as series regulators.[4] This has only recently become economically feasible, because several semiconductor manufacturers are now making power fet's, and the prices are beginning to come down.

There are several advantages to using a fet as a series regulator. The first is that the regulator has inherent current limiting, since the fet current is limited to $I_{dss}$. The second advantage is that "thermal runaway," as experienced with bipolar transistors, is not possible. The third reason is that because of the high impedance of the gate, essentially **no** gate current is required to control the fet current. This makes the fet gate even easier for the dc amplifier to control, and higher dc gain may be used.

The circuit shown in **fig. 9** uses a series power fet, a bipolar dc amplifier, and a second fet as a constant-current source. The constant-current source replaces the usual resistor load of the dc amplifier. By using the constant-current source as a load resistor, higher voltage gain can be obtained in the dc amplifier. The 5k adjustable source resistor allows the constant-current flow to be set to the value giving best temperature stability. This is usually around 0.33 mA. Because rather low current is used in the dc amplifier, a bipolar transistor displaying

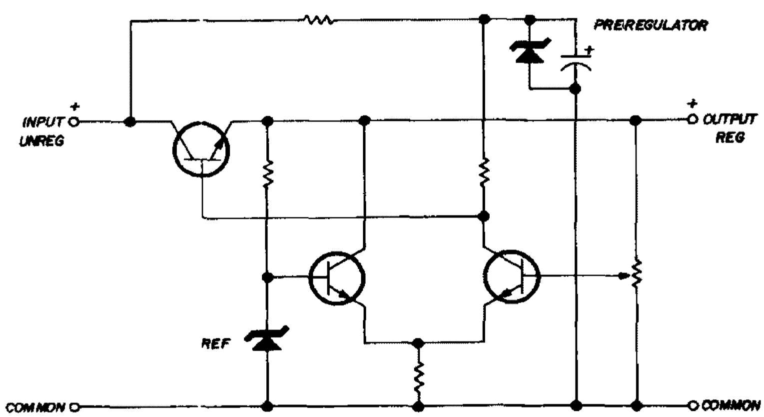



**fig. 7. Differential dc amplifier regulator. Zener reference runs at constant low current, improving reference stability.**

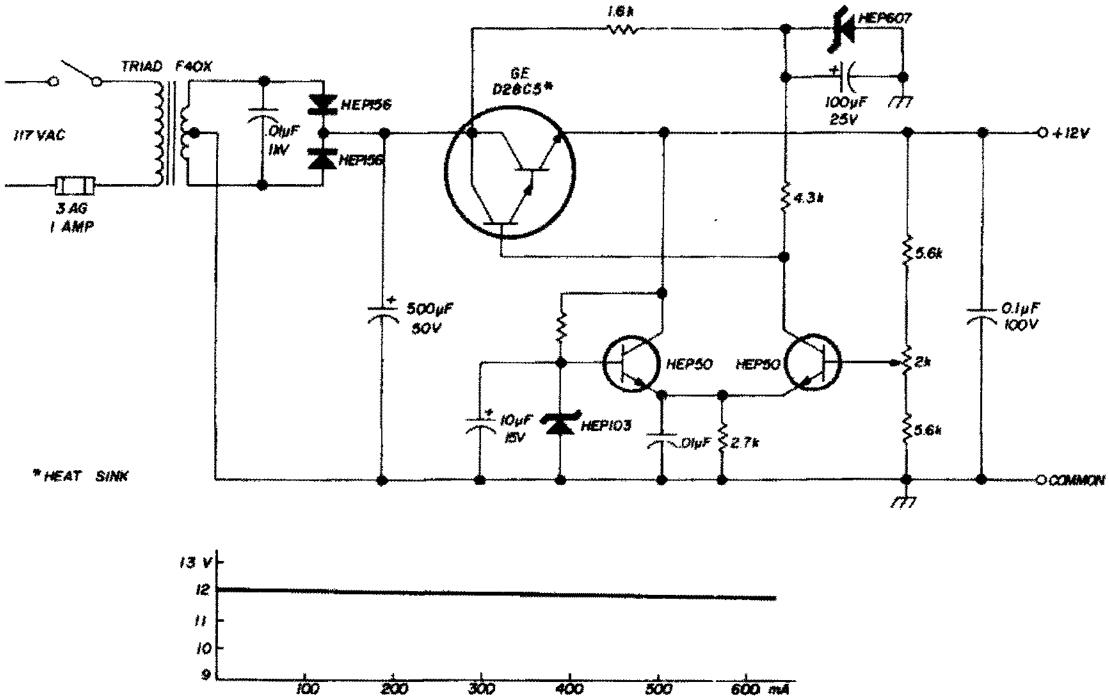



**fig. 8. Practical application of the dc differential amplifier.**

good $h_{fe}$ at low current (the Fairchild 2N3565) was used. Also, the zener was chosen to have a low impedance at 0.33 mA.

Since the supply of **fig. 9** uses choke input, a minimum load is required, which is supplied by the number 327 pilot lamp.

There are a couple of disadvantages to using an fet series regulator, but these are

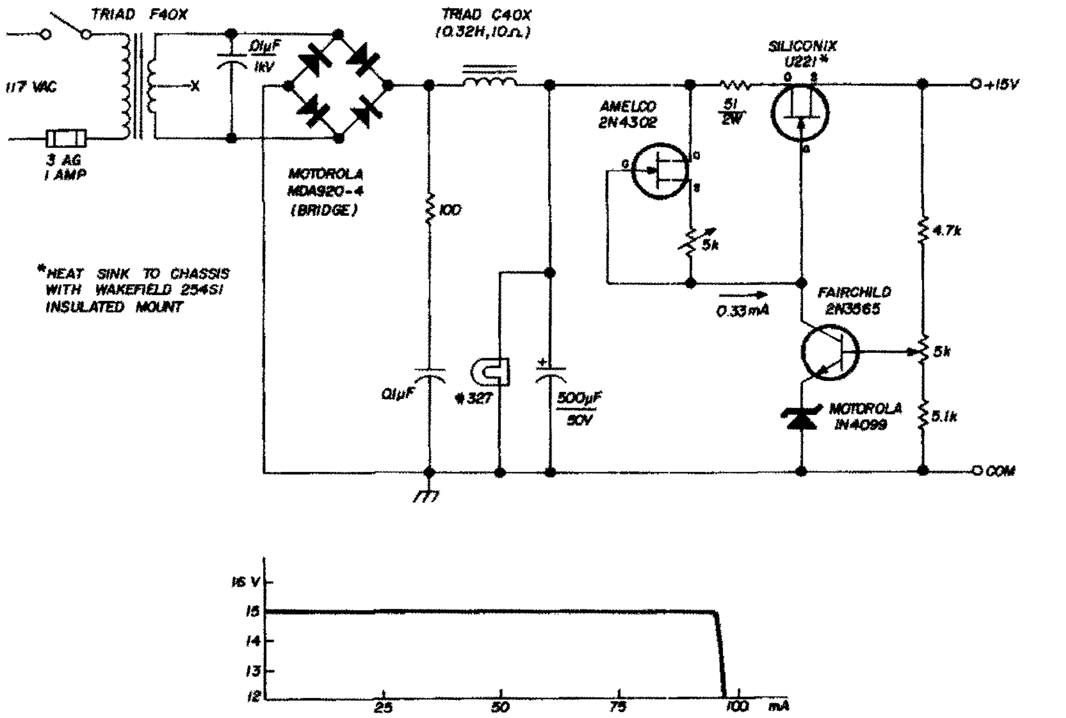



**fig. 9. Series-regulated supply using power fet's. The second fet replaces the usual resistor load as a constant-current source allowing higher voltage gain.**

not intrinsic faults. The obvious problem is cost; the less obvious one is loose specification of $I_{dss}$. The Siliconix U221 and U222 have $I_{dss}$ specifications of about 2 to 1, which are considerably tighter than those of the competition power fet's. As power fet technology advances, we can certainly expect to see prices drop and perhaps more predictable $I_{dss}$ from any given off-shelf device.

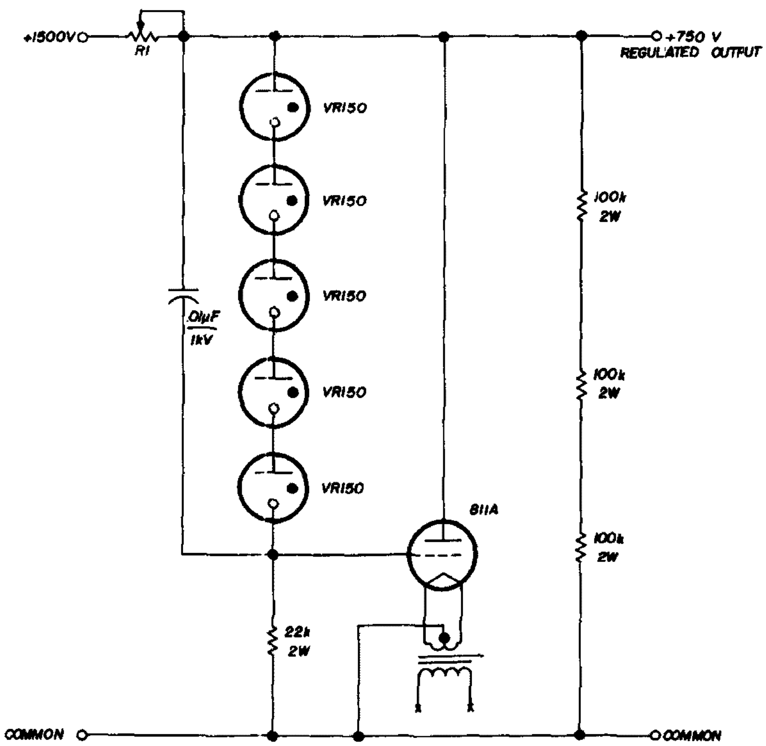



**fig. 10. Typical tube-type regulator. The 811A bias, controlled by the VR tubes, changes the voltage across R1 thus regulating the output voltage.**

## the shunt regulator

Before discussing the replacement of the discrete differential amplifier with an IC, there's one more circuit we should look at: the shunt regulator. The vacuum-tube shunt regulator has been in use by hams for many years. **Fig. 10** shows a typical HV type.[5] The operation is as follows:

If the +750 V output drops, the drop is directly coupled to the grid of the 811A by the VR tubes, dropping the bias. This causes the 811A to draw less current, thereby decreasing the voltage drop across R1, and raising the +750 V output.

A similar shunt regulator can be made using semiconductors; zeners replace VR tubes, and an enhancement-mode fet replaces the zero-bias triode. Such a direct equivalent circuit is shown in **fig. 11.** It would probably not be built today, because n-channel enhancement-mode fet's are not yet commonly available.

A more conventional approach to the shunt regulator is that of **fig. 12,** using a bipolar transistor. An actual circuit is shown in **fig. 13.** Another more complex shunt regulator is shown in **fig. 14,** using a differential amplifier. Note that both of these circuits

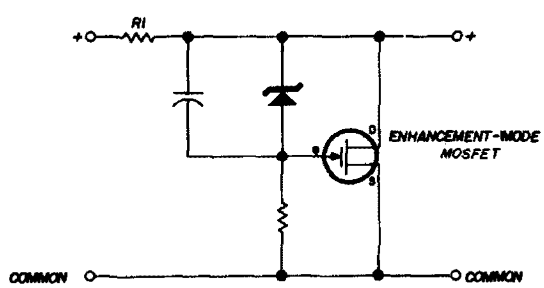



**fig. 11. Semiconductor equivalent of the vacuum-tube regulator. (N-channel enhancement-mode fet's are not yet commonly available.)**

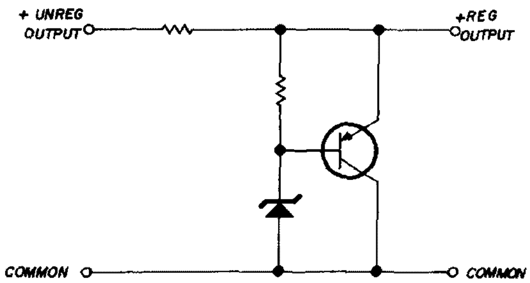



**fig. 12. Shunt regulator using bipolar transistor.**

(**figs. 13** and **14**) have a convenient feature: the collector (case on most power transistors) is grounded.

The shunt regulator is often overlooked because of its inefficiency when unloaded. However, in many systems the load is nearly constant, and the shunt regulator is used mainly to provide a low power supply source impedance. The regulator in **fig. 14** was designed to power a system consisting of about a dozen DTL integrated circuits, which had a nearly constant drain. Note also that the shunt regulator is ideal for the choke-input rectifier filter, because it provides a constant load.

## ic's as voltage regulators

We have seen two beginning steps toward integration of circuits. These are the "ref amp" and the "Darlington pair" as represented by General Electric's RA1 and D28C5. Each of these units contains two semiconductor devices. Modern IC technology has expanded greatly on the number of devices per package. It's now usual for IC's to have dozens, or even hundreds, of devices on one chip.

Among these complex IC's are a few designed specifically for voltage regulators. National Semiconductor makes the LM300, LM304 and LM305 voltage regulators; Fairchild makes the μA723 regulator; Motorola makes the MC1460G and MC1460R; and the Continental Devices Corporation makes the CMC513-4. Other companies[6,7] produce IC voltage regulators, but their prices are more than $10.00 and therefore aren't within the amateur price bracket in my opinion.

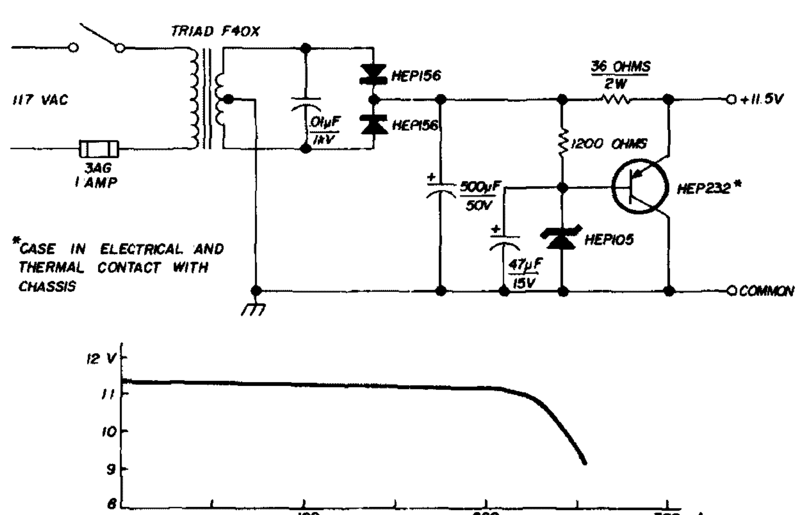
The LM300 was perhaps the first to hit the market and has been described in a previous article.[8] Its use in a +15 V supply is shown in **fig. 15**.

Since the original LM300 was first marketed, National Semiconductor has added the LM304 and the LM305. The LM305 is simply an improved LM300 (at a somewhat higher price); it can be plugged into the circuit of **fig. 15** to provide improved regulation. The LM304 is a **negative** voltage regulator; its use in a −15 V supply is shown in **fig. 16.**

Fairchild has entered the integrated regulator field with the μA723. The μA723 may be used either as a positive or negative regulator as shown in **figs. 17** and **18.** The many uses of the μA723 are covered in detail in reference **9.**

The least expensive IC regulator covered

**fig. 13. Practical shunt regulator supply using the bipolar transistor.**



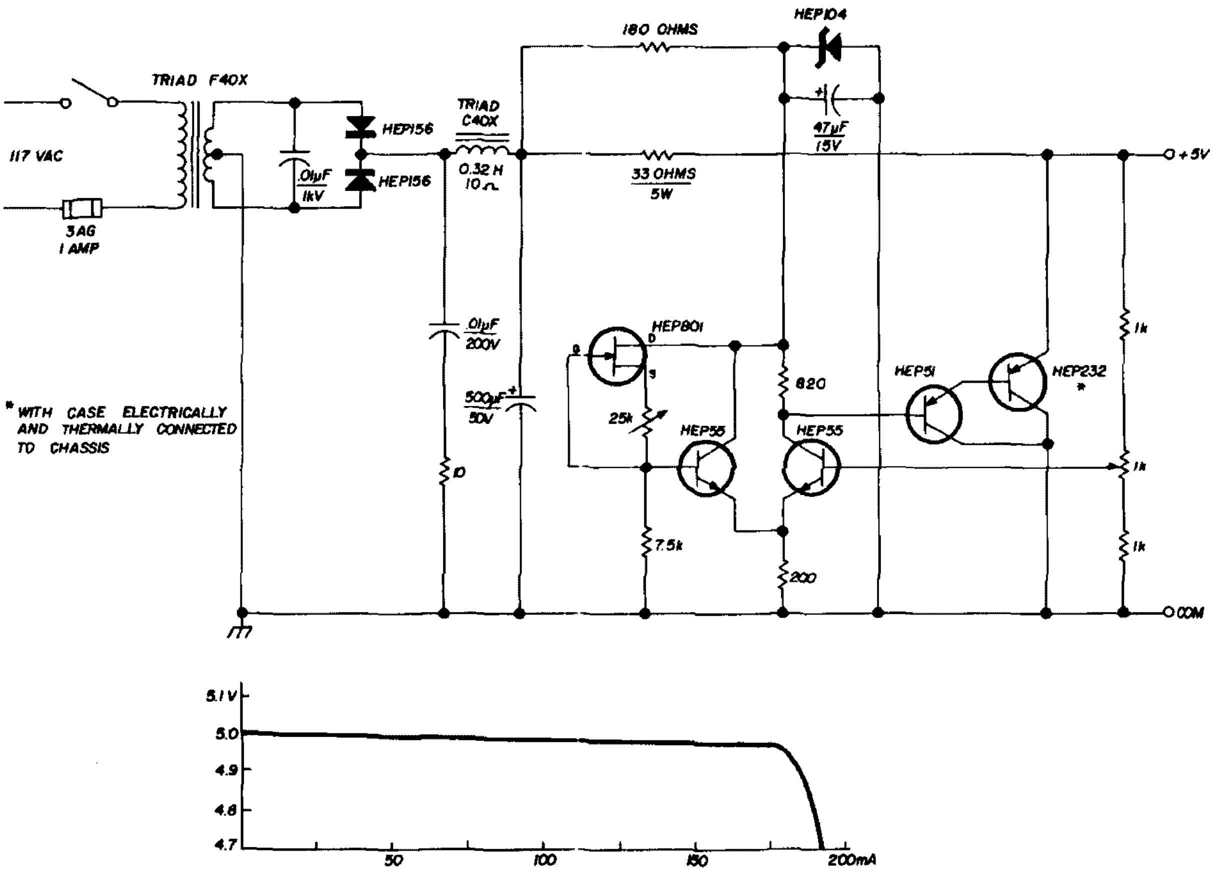
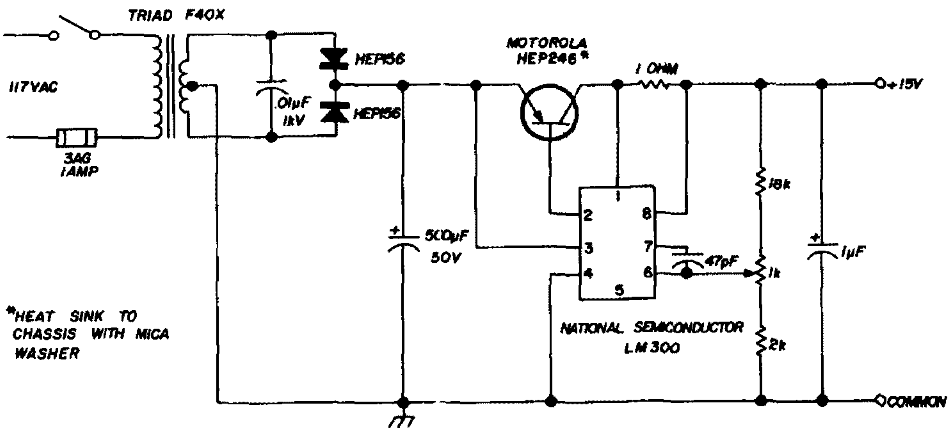



**fig. 14. Shunt regulator using a differential amplifier circuit. This was used to power a dozen IC's.**

here is a member of a family of five 15 volt regulators, each with the prefix CMC514. The CMC514-4 is the least sophisticated; it is also the only one in the ceramic-epoxy TO5 package (and therefore the least expensive). The CMC514-4 has only three leads, so it looks for all the world like a 2N3638—it isn't! **Fig. 19** shows it in a simple regulated supply; note that it has only an input, an output, and a ground lead. The CMC514-4 has a built-in thermistor, **fig. 20.** If too much current is drawn, the chip overheats and the termistor **turns the regulator off!** When chip temperature drops, the regulator begins regulating again.

A late entry into the regulator IC field is

**fig. 15. A regulated +15 V supply using an LM300 IC.**



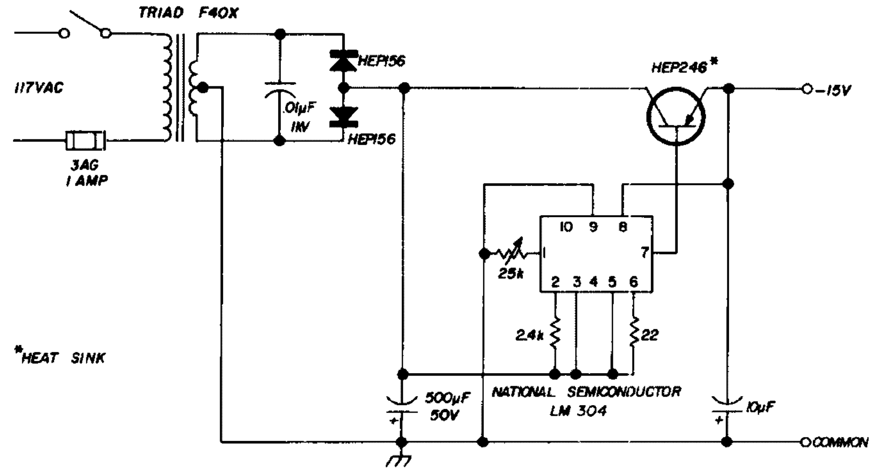



fig. 16. A −15 V supply. This IC is a negative voltage regulator.

the Motorola MC1460. It comes in a ten-lead TO-5 package (the MC 1460G) at $5.25 and in a small-diamond package (the MC1460R) at $6.75. The R model will, of course, handle more current, because it's easier to heat sink. In **fig. 21**, an MC1460G is used in a +15 V

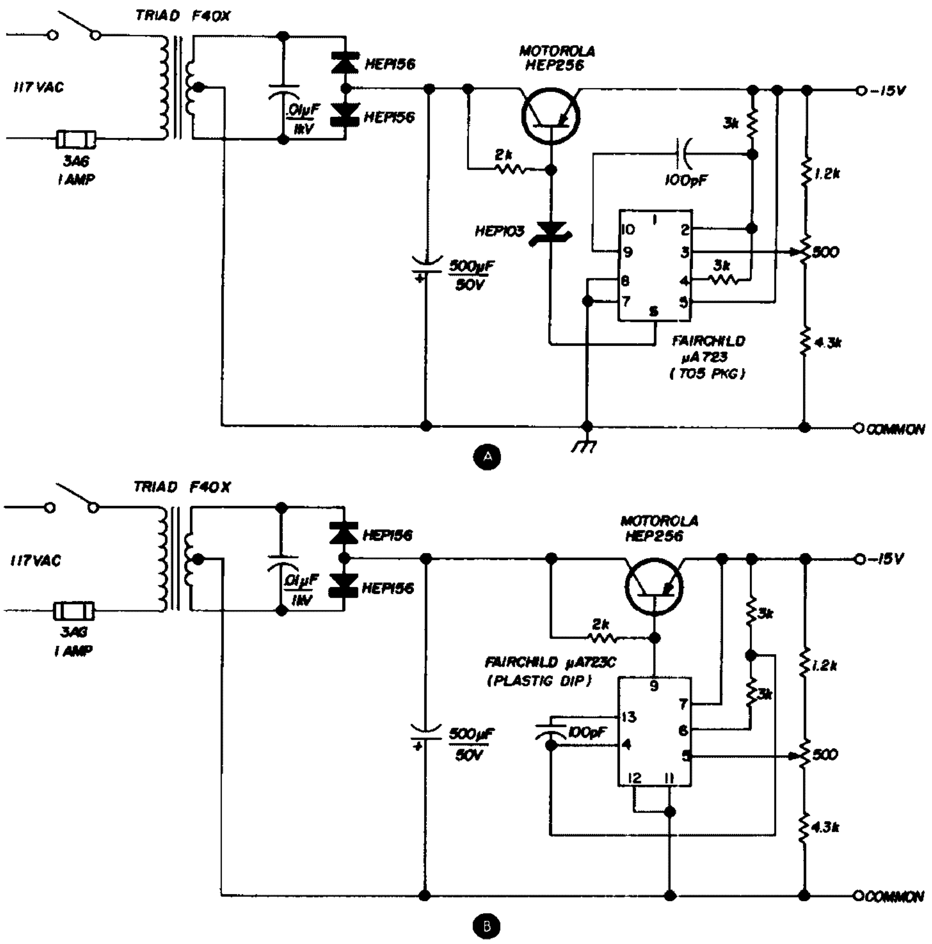



fig. 17. Another application of an IC in a regulated supply. Note that no external zener is needed in B.

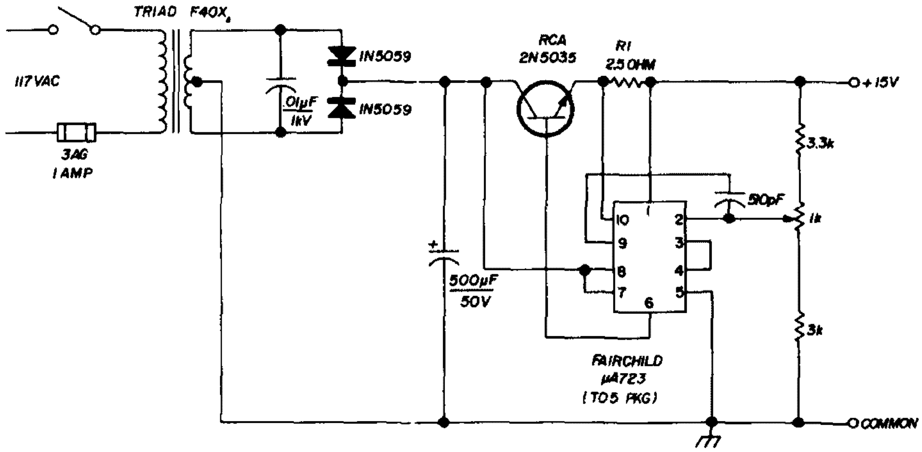
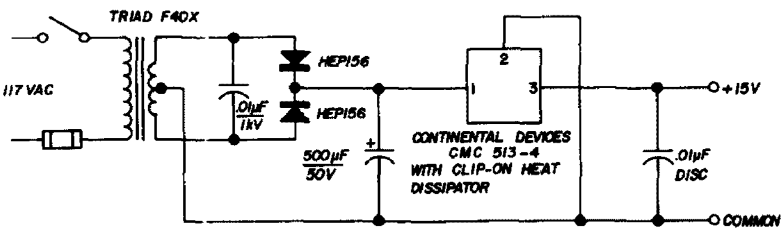



**fig. 18. This IC may be used either as a positive or negative voltage regulator; here it's shown in a +15 V regulated power supply. R1 determines current limiting. The value shown in for 200 mA.**

regulated supply. Details on how to use this IC are given in reference 10.

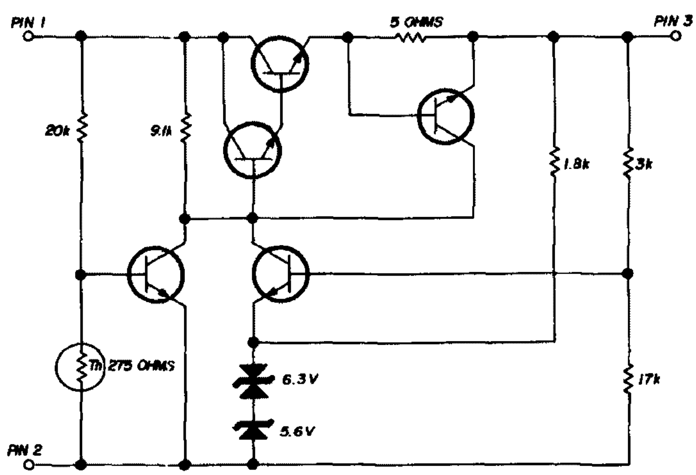
The MC1460 has a shut-down port (pin 2). In **fig. 21,** pin 2 is grounded, and so does nothing, but it can be used to turn off the regulator. If pin 2 is made high by a few volts, the regulator shuts down. This high state is low enough so that RTL, DTL or TTL logic levels can be easily used for control. This opens a myriad of possibilities: squelch control of the regulator to conserve power in the output stages of a receiver, shut-down upon demand by a temperature sensor, and so on.

**fig. 19. A +15 V regulated supply using the least-expensive IC. It comes in a ceramic-epoxy TO5 package.**




The MC1460 also has a built-in provision for current limiting. The 2.7-ohm resistor in **fig. 21** determines at what current the supply will limit. By making this resistor smaller,

**fig. 20. Internal circuit of the CMC 514-4 showing internal thermistor.**



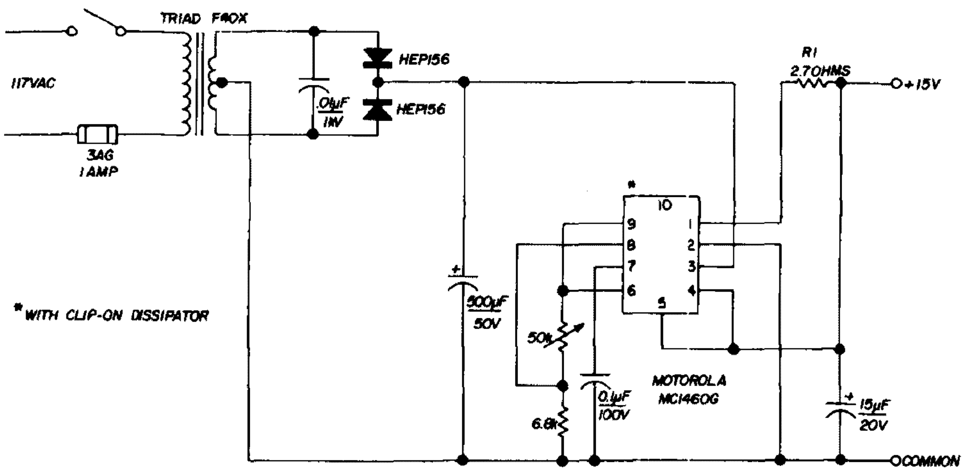



fig. 21. Motorola's MC1460G IC regulator in a +15 V supply has a current-limiting resistor (R1).

the regulator will limit at a larger current. A smaller resistor would probably not be used with the MC1406G, but could be used with the MC1460R.

## other ic's as regulators

We've seen how a number of special IC's can be used (for their intended purpose) as regulators. It's also possible to use others, not sold as regulators, for similar applications. Of course, nearly any operational amplifier can be used as the differential comparator in a regulator, providing that a reference, a series-pass transistor, and a fair number of other external parts are used. This makes a good regulator, but the complexity defeats one of the main reasons for using an IC—simplicity. Reference 11 shows how this is implemented using a $\mu$A709.

Two inexpensive audio amplifiers can be used as regulators. The General Electric PA234 and PA237 are 1- and 2-watt af output

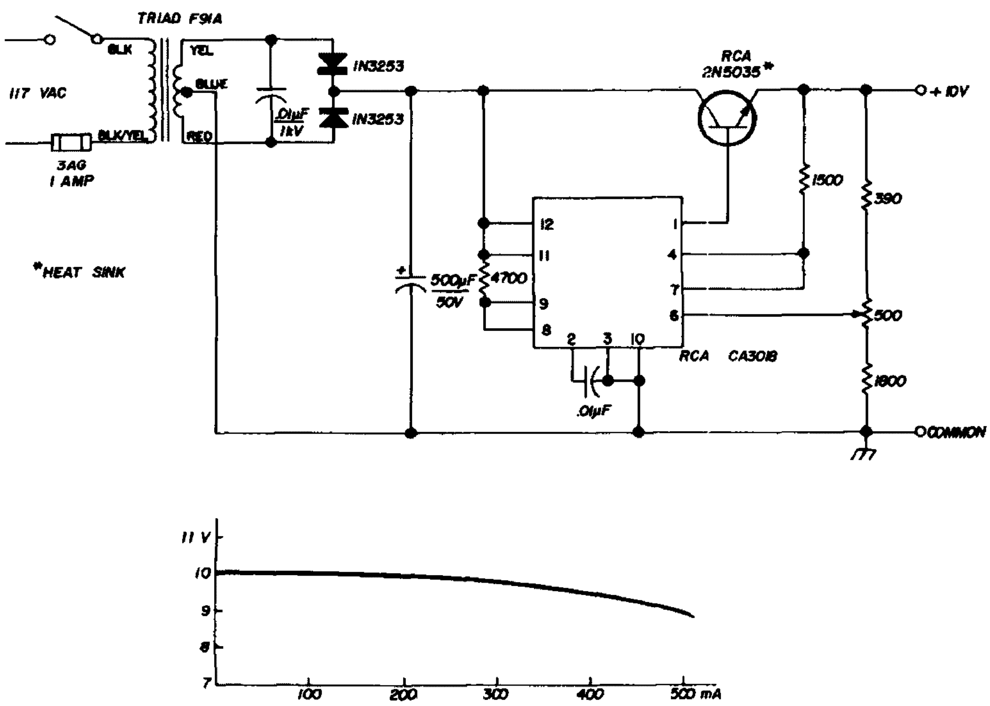



fig. 22. Another inexpensive IC in a voltage regulator circuit. An emitter-base diode in reverse breakdown provides the reference; an external series pass transistor must be added.

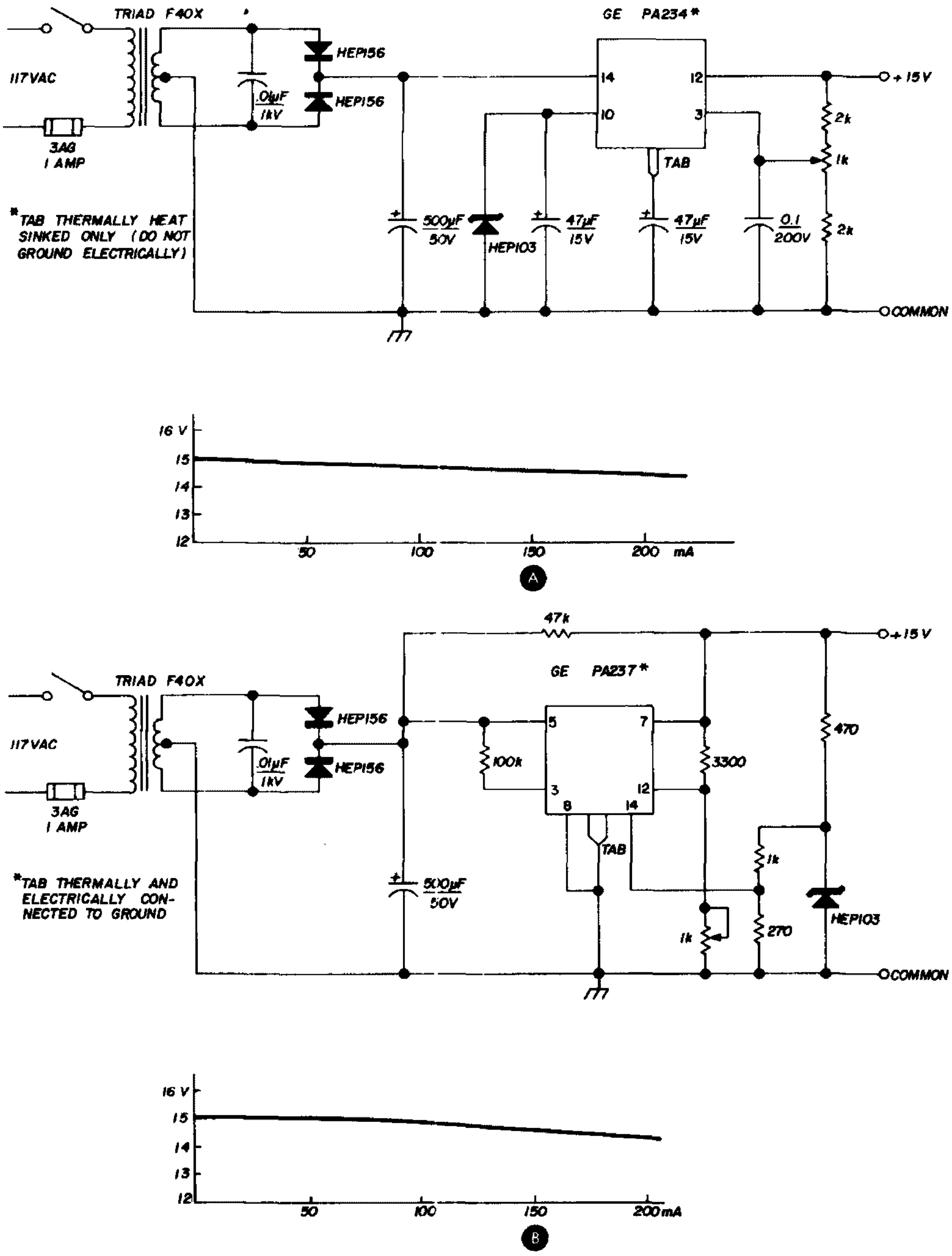


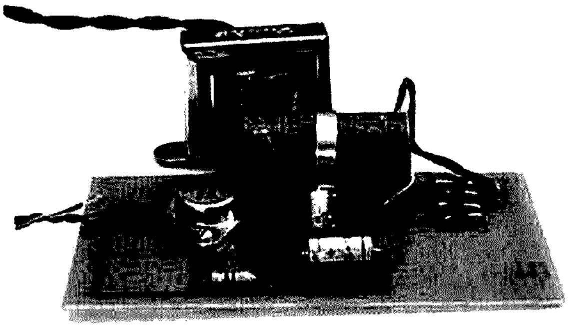

**fig. 23. Two inexpensive af amplifier IC's used as regulators in a +15 V supply. Current limitation is determined by package dissipation (0.8 W at 25° C).**

amplifiers costing $1.35 and $3.98 respectively. Their use in regulated supplies is shown in **fig. 23.** Their current limitation is set by the package dissipation: 0.8 watt at 25° C. This situation may improve with the introduction by G.E. of the new PA246, a five-watt audio amplifier.

The RCA CA3018 is another inexpensive IC that can be used as a regulator.[12] In this case, an internal reference is provided (by an emitter-base diode in reverse breakdown), but an external series pass transistor must be provided. This is shown in **fig. 22.**

As a final note to the use of IC's as regulators, it is strongly advised that care be used in layout. Since monolithic integrated circuit technology is based on the silicon epitaxial process, IC's almost always have active de-

vices capable of oscillation above 100 MHz. Indiscriminate use of long wire leads, capacitor bypassing, and similar practices will usually make a power supply that can be heard on the local fm set—if the parasitic doesn't eat up the IC first.

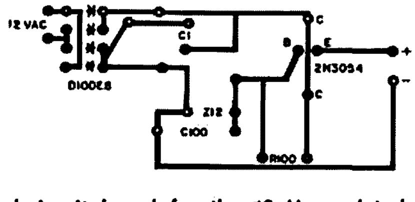


**Simple emitter-follower regulated 12 V supply for transistor circuits.**

## week end project

Most transistor circuits use 12 volts as their power source, and if you like to experiment with these devices (who doesn't these days?), here's a simple supply based on the emitter-follower principle.* It will deliver 12 volts at either positive or negative polarity, at a load current of 0.5 ampere. Regulation and ripple attenuation will satisfy most requirements. Its schematic is shown in **fig. 24.** It has ten components, which works out to about a dollar per component, and you can make the printed circuit board yourself.

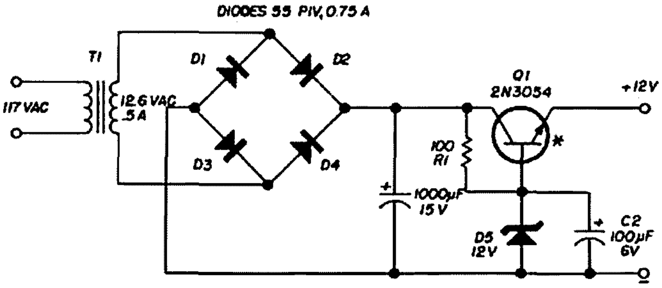
The photo of the PC board is the **positive** of the board. If you have a yen for photography, you can make a negative from which the board can be made. Another method is to trace the positive, turn the tracing paper over, and there's the mirror image. This is the actual diagram of the copper side of the printed board. Lay it out with Brady tape or acid resist, then etch away.




**Printed-circuit board for the 12 V regulated power supply shown in fig. 24. A full-size template is available from ham radio magazine for 25c.**

* The circuit of **fig. 24** and the photographs were submitted by M. J. Goldstein, VE3GFN.

For smaller load currents, a smaller transformer can be substituted. Be sure the transformer output voltage is close to 12 volts, because any difference between transformer output voltage and 12 volts will be dissipated across the transistor.




*MOUNTED ON HEAT SINK

**fig. 24. A +12 V regulated power supply using a bridge rectifier and emitter-follower regulator.**

### references


1. H. Olson, "Some Thoughts on Transistor Power Supplies," *73*, December, 1966, p. 10.
2. General Electric, "Reference Amplifier, Integrated Structures: RA1 A, B, C; RA2 A, B." Spec-Sheet 35.35, April, 1963.
3. Transitron, "Silicon References, Ref-Amp," Specification Sheet TE-1352E, November, 1968.
4. H. Wu, "Designing Power Supplies with FET's," *E.E.E.*, December, 1968, pp. 68-69.
5. ARRL, "The Radio Amateur's Handbook," 1962 edition, p. 234.
6. J. Stettner, "A Precision Regulated Supply Using WC-115 and WM-110," Westinghouse Integrated Application Note.
7. Beckman, "Model 802 Regulator Data Sheet 681223," Beckman Helipot.
8. H. Olson, "IC Regulated Supplies for Integrated Circuits," *ham radio*, March, 1968, p. 28.
9. Fairchild, "Application Notes, µA723 Precision Voltage Regulator," 20-BR-0064-78, 1968.
10. Motorola, "Specification Sheet, MC1460, MC1560," DS 9104, September, 1968.
11. Fairchild, "Linear Integrated Circuits Applications Handbook," Fairchild Semiconductor, 1967, p. 144.
12. L. Toth, "Low-Cost Voltage Regulator from One IC," *The Electronic Engineer*, October, 1968, p. 89.


**ham radio**

# increasing the reliability of warning lights

## Equipment warning lights must be reliable— here's how to increase their life

F. J. Case, W3NK, 2000 Kernan Drive, Baltimore, Maryland 21207

**When an indicator lamp tells** you something, it's important that the lamp be dependable. For example, if a dangerous situation is indicated by a red light, and a safe condition is indicated by that same light **not** being lit, you can imagine the problems that could arise if the indicator lamp bulb failed.

One of the criticisms of modern cars is the use of "idiot lights" to indicate critical items such as **OVERHEAT, NO OIL PRESSURE, DOORS NOT CLOSED,** and **BATTERY NOT CHARGING.** The logic is negative and not to be trusted. This accounts for the big sales of kits that return the car to the good old days of seeing what goes on by means of a direct readout instrument.

Let's extend this reasoning to indicator lamps used in electronic equipment. Most indicators are the **ON-OFF** type. This again points up possible dangers of depending on lamps to indicate a safe (**OFF**) condition. When I leave my station, I take a fast look around to make sure that no indicator lamps or dial lamps are lit. I don't trust the lamps; yet I must depend on them, but only after doing something to increase their reliability.

Incandescent-lamp life can be extended considerably by simply lowering the lamp terminal voltage. This will, of course, reduce the light output. However, for simple **ON/OFF** indications, you can spare a lot of light and still get a satisfactory indication.

Lamp life extension can be dramatic. For example, if you reduce the rated voltage to one-half, the lamp life will be extended to nearly 4100 times longer than if you applied the full rated voltage. Typically, a lamp with a guaranteed life of 1000 hours will operate for 4,096,000 hours when operated at half its rated voltage.

Where does this magic number 4096 come from? It's based on the relationship of the ratio of the rated voltage to the derated voltage raised to the twelfth power, or:

$$\text{Life factor} = \left(\frac{\text{Rated Voltage}}{\text{Derated Voltage}}\right)^{12}$$

### graphical solution

You can do the arithmetic the hard way and multiply the ratio by itself 12 times, or you can solve the problem with logarithms. Another way is to cut out **fig. 1** or (preferably) make a Xerox copy. Or you can make your own graph on a piece of five-cycle semi-log paper.

Fig. 1 shows that lamps can be as bright as you wish; lamp light will depend on the derating voltage ratio.

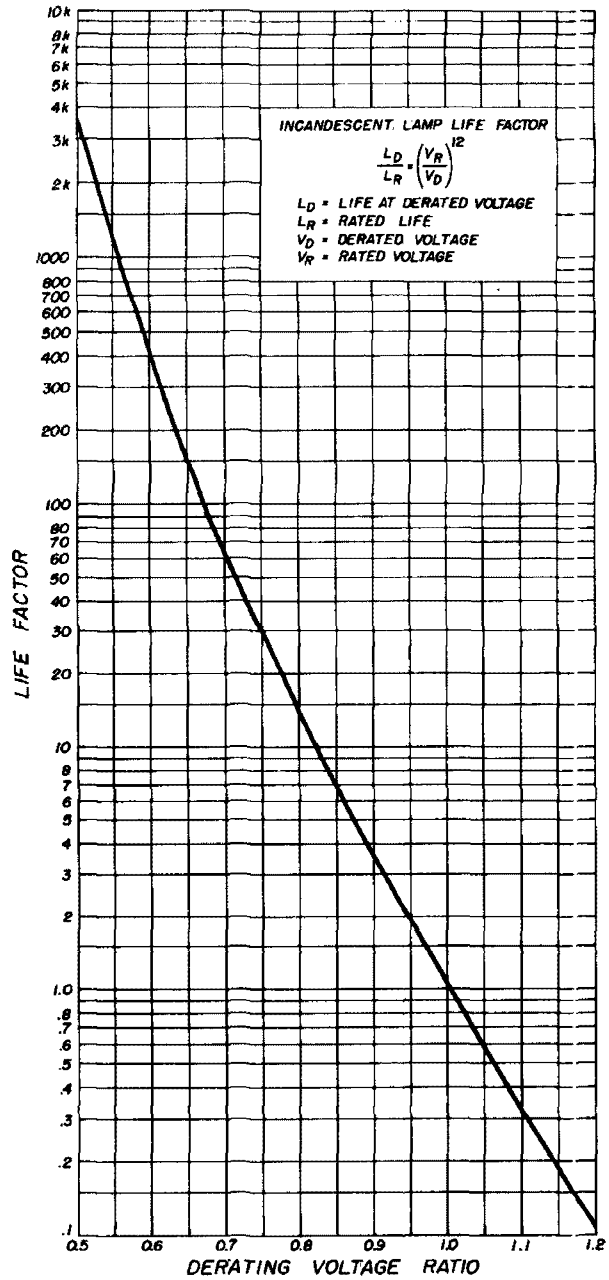



**fig. 1. Effect on pilot-lap life when a derating voltage ratio is applied. A small sacrifice in brilliance means a drastic increase in reliability.**

## an example

About five years ago, I got disgusted replacing Christmas-tree lamps and decided to do something about it. I connected a 25 V, 2A filament transformer in series bucking to provide 90 volts, which was applied to the lamps. The voltage drop made a barely perceptible difference in brightness, but I haven't had to buy a bulb in five years.

Let's see what the 90 volts did for me. The rated voltage (for the series string) is 115 volts. I applied 90 volts. The ratio is 115/90, or 1.277. By the formula above, the life factor will be $1.277^{12}$ or 19. So, whatever the rated life of these cheapies, they will last 19 times longer than when operated at 115 volts.

## indicator lamp circuits

I modified the indicator lamp circuits in my ham equipment by adding series resistors. For type 328 lamps, I added a 22 ohm, 1/2 W resistor, which reduced the voltage from 6.3 to 3.15 V. Brightness is more than adequate; in fact, I'm tempted to reduce the voltage even further. The ratio 6.3/3.15 equals 2. When raised to the 12th power, this gives a life factor of 4096, or more than 4 million hours. This works out to about 467 years, so I think I'll let well enough alone.

For the type 47 lamps, a 15 ohm, 1 W resistor results in a terminal voltage of 4 V. The ratio is 6.3/4, which is a life factor of 233—still pretty good.

As mentioned earlier, the light intensity is reduced, but the life factor goes up much more rapidly than the intensity goes down.

The type 328 lamp gave more than adequate light at half voltage. The type 47 lamp, on the other hand, needed 4 volts to provide a comparable brightness. So do a little experimenting. Get the brightness you need, then do a little arithmetic or use **fig. 1** to project the lamp life. Remember too, that overvoltage will shorten the life of your lamps just as dramatically.

In addition to peace of mind, look at all the money you'll save. In 100 years it will be enough to get that new rig.

### bibliography

"Photomods Brochure," Clairex Electronics.

**ham radio**

# build your own tilt-over antenna mast

Here's the answer to a structurally sound, low-cost mast using ordinary hand tools and readily available hardware

Sidney Wald, W6KRT, 6430 Ellenview Avenue, Canoga Park, California 91304

**This tilt-over antenna mast** fulfills the need for a lightweight, low-cost amateur rotary beam support that can be raised or lowered in a few minutes by one man without assistance. As shown in the accompanying sketches and photographs, the design requires materials and tools commonly available to most amateurs, takes up only modest backyard space, and unlike lattice-type metal or wood towers, is quite unobtrusive, especially when painted to blend with the background.

The project originated when I was pondering ways and means for supporting a lightweight Mosley TA33 Jr. three-element beam. The obvious solution to the problem would be to purchase and erect one of a large number of available steel towers, but this was ruled out for esthetic and financial reasons. Using the rooftop or chimney was likewise unacceptable. One alternative remained; a new design had to be conceived that was structurally sound, low in cost, and easy to build and manipulate.

Over the weeks many ideas were tried and rejected, and during the construction period many revisions were made. The tilt-over mast and antenna have been in operation for several months, during which they've been lowered and raised many times without incident. I've included a detailed stress analysis, complete with quantitative data in this article (see Appendix). The analysis considers loads encountered in raising and lowering the structure, as well as those resulting from wind velocities up to 80 miles per hour.

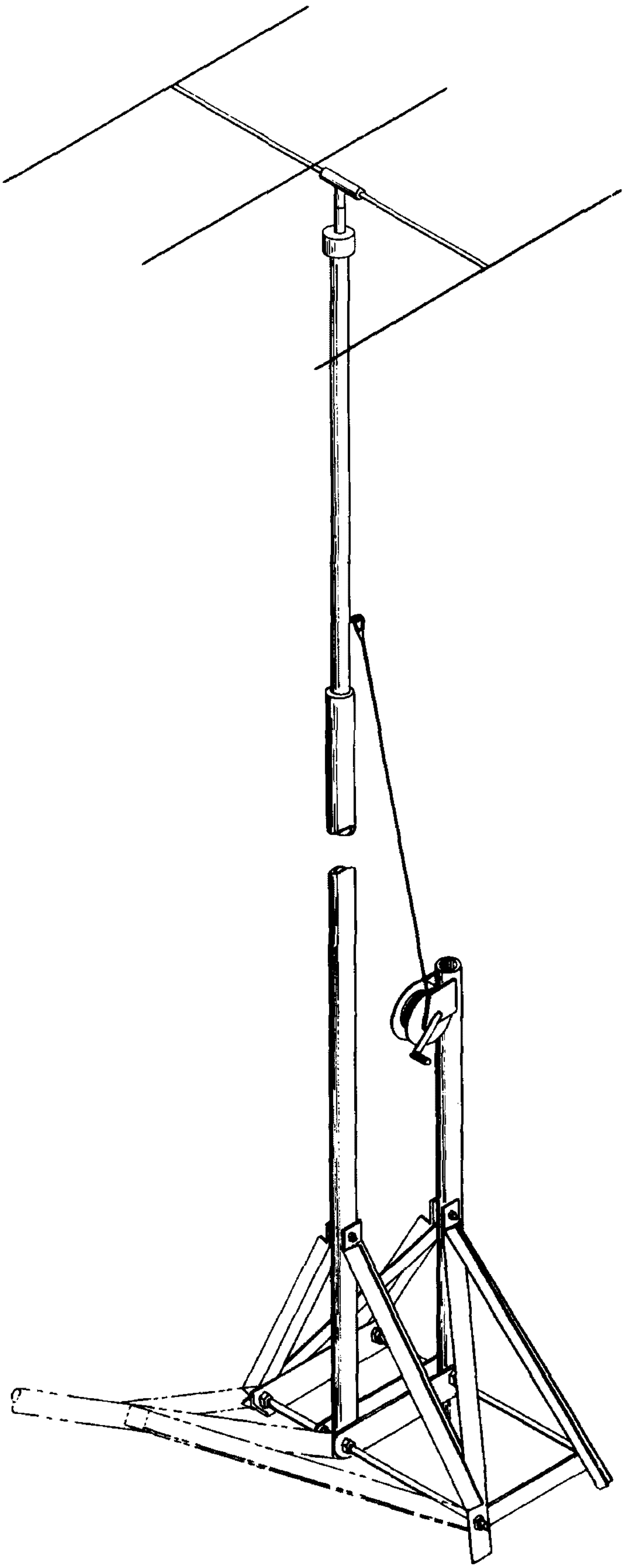


**fig. 1. Complete tilt-over mast structure showing mast-hinge assembly.**

Four novel features are included in the design and construction of the mast:

A boat winch mounted on a short, fixed support pipe for raising and lowering the mast.

A rigid hinge or pivot assembly at the base, which prevents lateral deflection of the mast and antenna assembly during erection.

Two sets of symmetrically placed guying cables to maintain the mast in the vertical position without buckling.

Auger or screw anchors in the soil as high-strength fastenings for the guying cables and winch pipe support.

In all cases, safety factors of at least 2 to 1 were used to compensate for any small design approximations and to insure adequate resistance to high winds.

## construction

The first step in the project is to accumulate the parts listed in **table 1.** The only precaution here is to be sure to purchase 1/8-inch diameter, 1800-pound-breaking-strength stranded-steel aircraft cable and **not** galvanized *iron* sashcord (which resembles the proper material) with a breaking strength of only 540 pounds. Stainless steel cable is even better, but it's quite expensive.

The screw anchors may be obtained from Spaulding Products, Frankfort, Indiana. Other materials are obtainable from plumbing and hardware supply sources. The winch is obtainable from Sears-Roebuck at about $6.00 for the smallest unit, with a capacity of 1000 pounds. Good construction practice for any device that must withstand weather calls for at least two coats of paint. Don't overlook this essential item.

## ground work

Selection of a site comes next. Be sure that an unobstructed radius equal to the mast, plus antenna and rotor height, is available in the vertical plane for raising and lowering. In the example shown here,

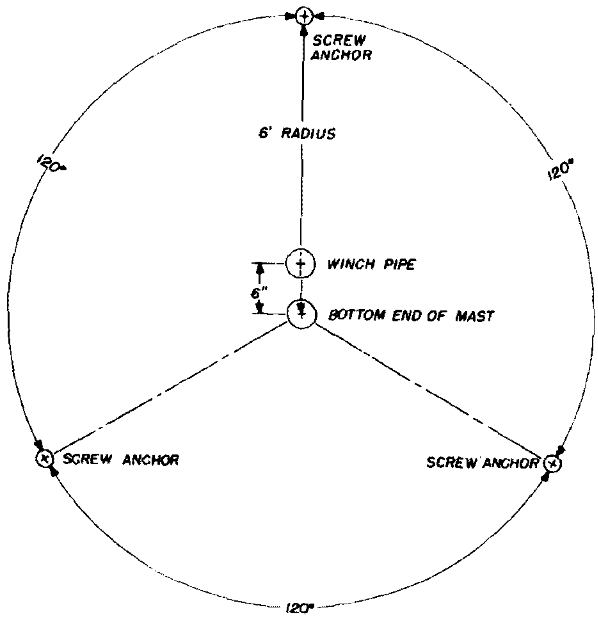



**fig. 2. Plan view of the tilt-over mast base site.**

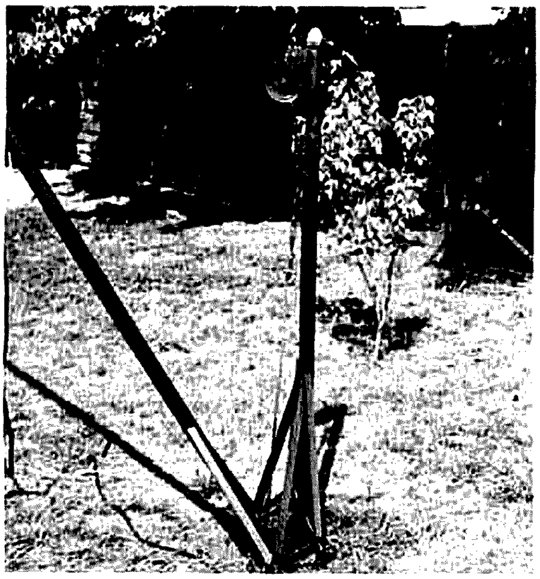
this amounts to a distance of about 35 feet plus adequate clearance for the antenna boom and elements.

After you select the location of the mast base and lay out the site as shown in **fig. 1,** drive the 2-inch pipe until it projects about 2 inches above the ground. This forms the socket that receives the 1½-inch winch pipe. (The location of this socket is about 6 inches to the rear of the bottom end of the antenna mast.) Then insert the 1½-inch winch pipe into the 2-inch pipe until the winch pipe bottoms. Fasten the assembly together with a 3/8-inch steel bolt, lock washer and nut. (See **fig. 2.**)

Next, drive the three screw anchors into the ground 120 degrees apart at a radius of 6 feet from the bottom end of the antenna mast as a center (see **fig. 1**). The technique for installing these anchors is as follows: After the location is spotted, make a shallow depression 6 inches in diameter by 6 inches deep. Then using a length of 3/4-inch pipe as a handle through the eye of the screw anchor, twist the auger end of the anchor into the ground, using downward pressure while

**table 1. Parts list for the tilt-over mast.**

| part | description | quantity |
|---|---|---|
| galvanized iron pipe | 2-ft x 2-inch diameter (nom) | 1 |
| galvanized iron pipe | 7-ft x 1½-inch diameter | 1 |
| galvanized iron pipe | 10-ft x 1½-inch diameter | 1 |
| galvanized iron pipe | 21-ft x 1¼-inch diameter | 1 |
| extruded aluminum alloy pipe | 3-ft x 1¼-inch diameter | 1 |
| boat winch | 1000-lb capacity | 1 (Sears) |
| aircraft cable | galvanized steel<br>1/8-inch diameter<br>1800 lb rating | 210 ft |
| turnbuckle | 5/16-inch x 8-inch | 7 |
| screw anchor | 4-ft x 6-inch diameter helix | 3 (Spaulding Products) |
| wire rope clip | 1/8-inch | 58 |
| wire rope thimble | 1/8-inch | 39 |
| galvanized or stainless steel bolts | 3/8-inch diameter x 4-inch long | 5 |
| eye bolt, galvanized iron | 3/8-inch diameter x 4-inch long | 1 |
| guy ring | To fit 1¼-inch pipe | 1 |
| threaded galvanized steel rod | ½-inch diameter x 2 ft | 2 |
| galvanized angle iron | 1¼-inch flange x 1/8-inch thick | 15 ft |
| miscellaneous locknuts, lock washers and flat washers | | 30-40 |
| paint | | ½ gallon |

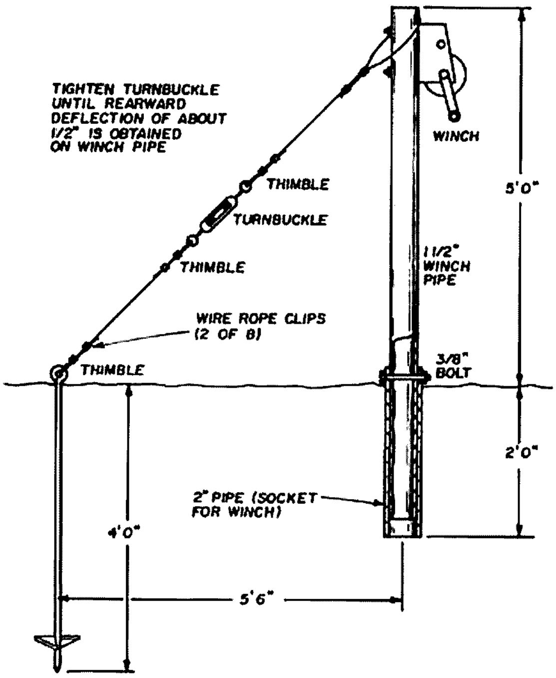


turning. Initial moistening of the hole may be necessary to start the anchor.

Twist the anchors into the ground until only the circular eyes protrude. When installed in average backyard soil, these anchors can withstand a pull of over 2000 pounds.

Mount the winch near the top of the $1\frac{1}{2}$-inch pipe at a height of about 5 feet above the ground, using two 3/8-inch diameter steel bolts, lock washers and nuts.

**fig. 3. Winch pipe support.**




The top of this pipe is then anchored to the rearmost screw anchor eye by a tension assembly consisting of steel cable, turnbuckle, thimbles and wire rope clips as shown in **fig. 2.**

## pivot assembly

The next step is the construction of the pivot, or hinge assembly about which the mast tilts. The basic materials here are

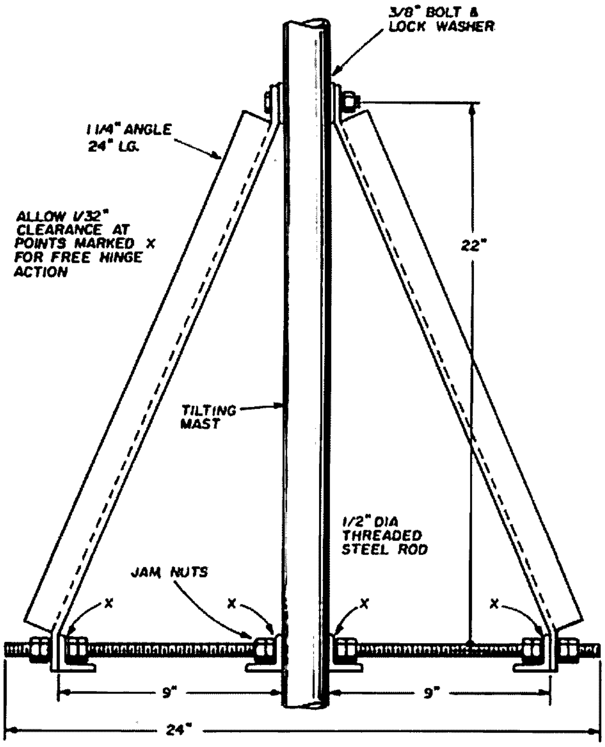



**fig. 4. Mast hinge assembly detail. Leave nuts slightly loose to permit antenna mast to pivot; use jam nuts (two on each side) to secure.**

various lengths of angle iron to form the rigid triangular gussets and the threaded $\frac{1}{2}$-inch diameter steel rods that form the transverse axle and compression members as shown in **figs.** 3 and **4.**

The flanges on angle-iron members are trimmed and bent at the ends to permit them to be joined as shown. When properly assembled, the hinge structure permits the mast to tilt about the pivot axis, while the axis is rigidly held in space by the balance of the triangular hinge structure. The latter is stationary with respect to the winch pipe. Tighten all nuts securely with

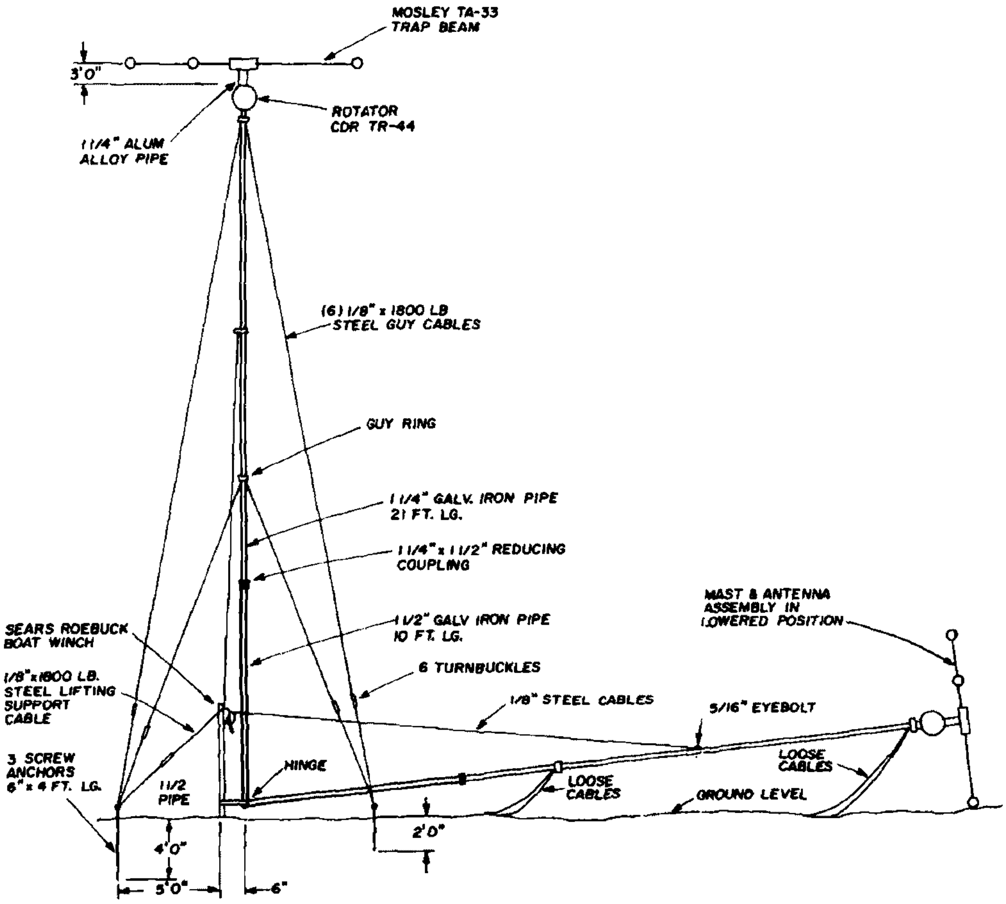



**fig. 5. Completed tilt-over mast.**

the exception of those on the pivot axis (**fig. 3**). These are left slightly loose to permit the triangular assembly on the antenna mast to rotate about the axis of the threaded rod. Be sure to use flat washers and lock washers, or locknuts.

The exact dimensions of the hinge assembly are not critical, and the sketches and photographs illustrate construction techniques. In general, the distance between mast and winch pipe should be kept short (about 6 to 8 inches). The triangular assembly of **fig. 5** shows dimensions that will give adequate lateral support for the 35-foot mast.

## mast assembly

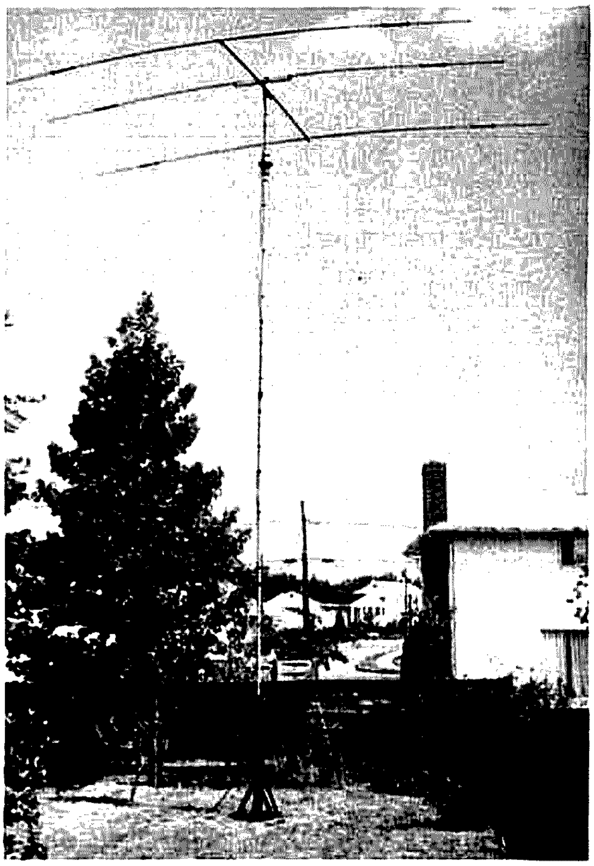
The next step is to lay the 1½-inch pipe on the ground and assemble it to the movable triangular hinged member via the two drilled holes, which receive the threaded rod and upper 3/8-inch bolt respectively. The assembly at this point should resemble **fig. 6.** Now screw the reducing coupling to the free end of the 1½-inch mast pipe, and then follow this with the 1¼-inch pipe.

You should now have about 31 feet of iron pipe on the ground. Drill a 3/8-inch hole 23 feet from the bottom end of the 1½-inch pipe. This receives the 3/8-inch eyebolt, which will be the main lifting member when the mast is raised. (See

**table 2. Parameters used in raising-and-lowering stress analysis.**

| part | lifting stress (lbs) | breaking strength (lbs) |
|---|---|---|
| main lift cable | 539 | 1800 |
| eyebolt | 117 | 2000 |
| winch pipe | 117 | — |
| winch | 539 | 1000 (rated load) |
| winch support cable | 737 | 1800 |

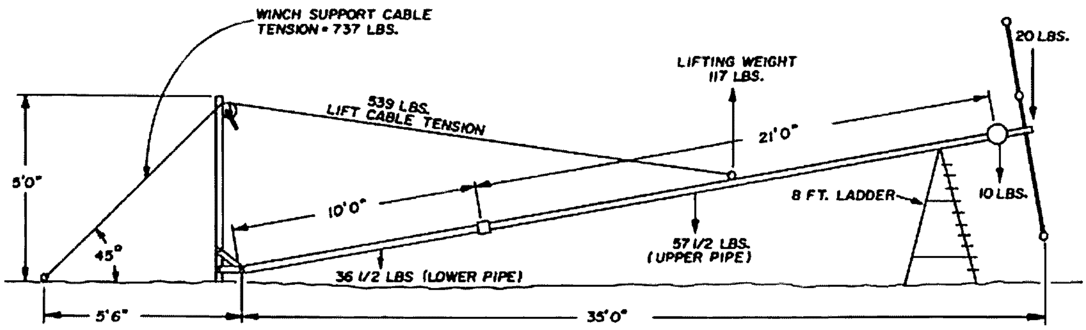
photograph.) Following this, attach the guy ring at the midpoint (16½-foot) level, and fasten three guy cables to the ring using thimbles and wire-rope clips. Mount the antenna rotator at the end of the mast pipe assembly, and fasten three guy cables to the base of the rotator as shown in the photograph.



## up she goes

At this time you're ready to raise the mast and rotator to about 8 feet, so that the beam can be assembled to the rotator. To do this, attach one end of a length of 1/8-inch aircraft cable to the 3/8-inch eyebolt, and attach the other end to the drum of the winch, allowing one or two layers of cable to accumulate on the drum before taking up tension.

If the winch is now cranked until the rotator clears the ground by 9 or 10 feet, an 8-foot ladder may be placed under the mast near the rotator, and the winch can be cranked backward until the mast is supported in part by the ladder.




**Clockwise moments** $= (\text{beam wt} \times 35) + (\text{rotor wt} \times 31) + (\text{wt of lower pipe} \times \frac{10}{2}) + (\text{wt of upper pipe} \times 20\frac{1}{2})$

$= (20 \times 35) + (20 \times 31) + (36.5 \times 5) + (57.5 \times 20.5)$
$= 2683$ **ft lb**

**(lifting wt at eyebolt,** $W_L$**)** $\times 23$ **ft** $= 2683$ **ft lb**

$W_L = 2683/23 = 117$ **lb**

**Angle A** $= \arctan 5/23 = 12.3°$

**Cable tension** $= \dfrac{117}{\tan 12.3°} = 539$ **lb**

**Stress on winch pipe in compression is equal to** $W_L = 117$ **lb**

**Tension on winch support cable** $= \dfrac{(539 \times \cos 12°)}{\sin 45°} = 737$ **lb**

**Stress on screw anchor** $= 737$ **lb**

**fig. 6. Stress analysis for lifting operation.**

The antenna assembly is now secured to the 3-foot length of $1^{1}/_{4}$-inch extruded thick-wall aluminum alloy pipe, and the pipe is inserted into the rotator and firmly clamped into place. Since the TA33 Jr. beam weighs only 20 pounds, this operation is not too difficult.

The rest of the operation consists of raising the entire assembly gradually with the winch until almost vertical, attaching the two sets of guy cables to the screw anchors and bringing the mast into true vertical position using the turnbuckles for adjustment.

The tension in the lifting cable may now be reduced, since its function has been fulfilled, and it will not be used again until you want to lower the antenna mast.

The middle set of guys prevents the slender pipe structure from buckling while the main guy cables (attached to the rotator) maintain the mast in a vertical position. True vertical is established near the base, using a carpenter's level, while the upper part of the mast is visually aligned using neighboring structures as references. You can now stand back and proudly view your mast and antenna, which will appear as in **fig. 7**. Give them a coat of paint, and you're ready for many months of trouble-free operation.

**table 3. Wind load stress analysis parameters.**

| part | wind load stress | load carrying ability |
|---|---|---|
| guy cable | 870 lbs | 1800 (breaking strength) |
| screw anchor (in dense clay soil) | 870 lbs | 2000 (holding strength) |
| rotator casting anchorage | 7000 psi | 14000-23000 psi (yield strength) |

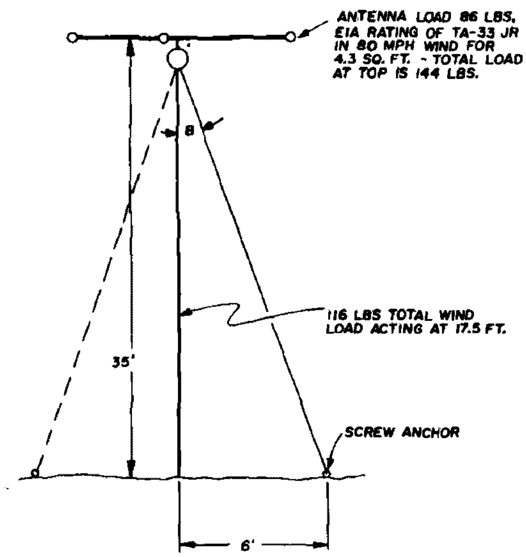



**fig. 7. Wind load stress analysis.**

**Angle B = arctan 6/35 = 9.7°**

**Stress in projected area of mast pipe = approximately**

$$\frac{(2 \text{ in} \times 420 \text{ in})}{144} = 5.83 \text{ sq ft}$$

**Mast wind load** $= \frac{5.83}{4.3} \times 86 = 116$ **lb acting at half-way point**

**This is equivalent to an additional wind load of 116/2 = 58 lb acting at the top**

**Thus, total wind load at top = 86 + 59 = 144 lb**

**Assuming cable prestress of 25 lb, cable tension =**

$$25 + \frac{144}{\tan 9.7^\circ} = 870 \text{ lb}$$

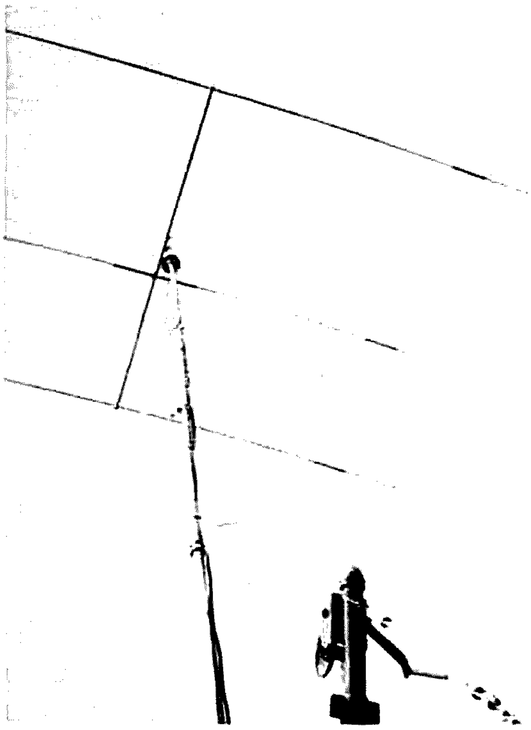
## appendix

### stress consideration

Two separate structural stress situations were taken into account in the design of the tilt-over mast. The first occurs only during raising and lowering, while the second is concerned with resistance of the structure to high winds.

In the first situation the critical design parameters are:

1. Lifting cable tension
2. Lifting eyebolt strength
3. Deflection of the mast pipe assembly
4. Stress on the winch, winch support pipe and winch support cable
5. Tension on the screw anchor holding the winch support cable



Referring to **fig. 8** a summary of these stresses compared to design breaking strength is shown in **table 2**.

Deflection of the mast pipe during lifting was a maximum of about 12 inches, which is well within safe limits for the 32-foot length of iron pipe. Once the assembly is in the vertical position, all stresses in the lifting components are relieved, with the exception of the winch-support cable, which may be maintained at some nominal tension (say 50 pounds) to keep the support pipe rigidly fixed.

Referring to **fig. 9**, the cable stress is 870 pounds compared to a breaking strength of 1800 pounds and a screw anchor holding ability of 2000 pounds. The cable-fastening point on the lower aluminum rotator casting will be subjected to a stress of $\frac{870}{0.125} = 7000$ pounds per square inch, compared to a yield strength of 14,000 to 23,000 pounds per square inch. **Table 3** gives the data.

It should be noted that all wind load stresses are based upon the Mosley TA33-Jr. EIA rating of 86 pounds for a projected area of 4.3 square feet at 80 miles per hour wind velocity.

**ham radio**

# a different approach to amplitude modulation

Expensive modulator components are eliminated in this design— modulation is provided by the rf amplifier power supply

Courtney Hall, WA5SNZ, 7716 LaVerdura Drive, Dallas, Texas 75240

**A characteristic of conventional** a-m transmitters is their bulky (and expensive) modulator, modulation transformer and power supply. How would you like an a-m transmitter capable of 100-percent modulation that doesn't require these components? The circuit described in this article may be the answer.

Although the system imposes a slight sacrifice in efficiency, this is compensated for by lower cost of original equipment, less weight, and smaller size compared to conventional a-m designs. Fewer components also mean increased reliability. This system uses a variable-current generator to modulate the output of a regulated power supply in accordance with the audio signal impressed on its input.

## standard a-m system

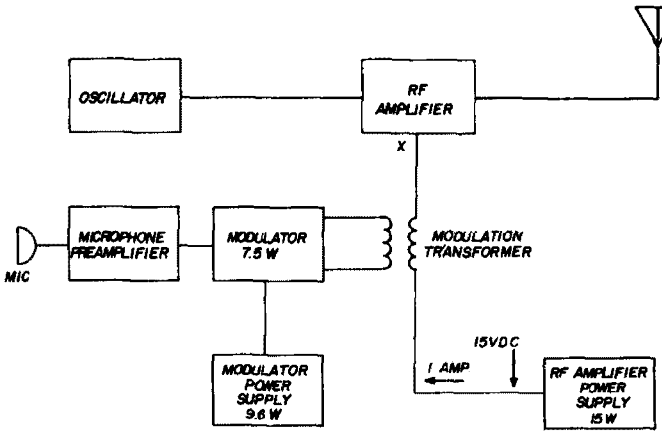
A block diagram of a conventional a-m transmitter is shown in **fig. 1.** Typical waveforms, voltage, current and power are shown. Oscillator and microphone preamplifier power supplies have been omitted for simplicity.

With no microphone excitation, the carrier signal is unmodulated. The dc input to the rf amplifier is 15 watts. For 100-percent modulation, 7.5 watts of audio power are required from the modulator. Point X (**fig. 1**) then increases from zero to 30 volts.

Assuming class B modulation, 9.6 watts are required from the modulator power supply. The modulation transformer must be capable of handling the power in a linear

fashion. Since this transformer is costly, bulky and heavy (except in very low-power transmitters), it would be advantageous to eliminate it.

**fig. 1. Conventional a-m transmitter with power requirements for 100-percent modulation.**




Additional savings in cost and size can be gained by eliminating the modulator. The circuit to be described requires no modulation transformer or modulator, but performs the same function by using the rf amplifier power supply regulator as a *combination* dc regulator/modulator.

## circuit description

**Fig. 2** is a schematic of a conventional regulated power supply that could be used to supply power to the rf amplifier. A sample of the output voltage (point A) is compared with the reference-diode voltage by a differential amplifier. The error signal at the differential amplifier output drives the series transistors, so that the output voltage is held constant as line voltage and load conditions vary.

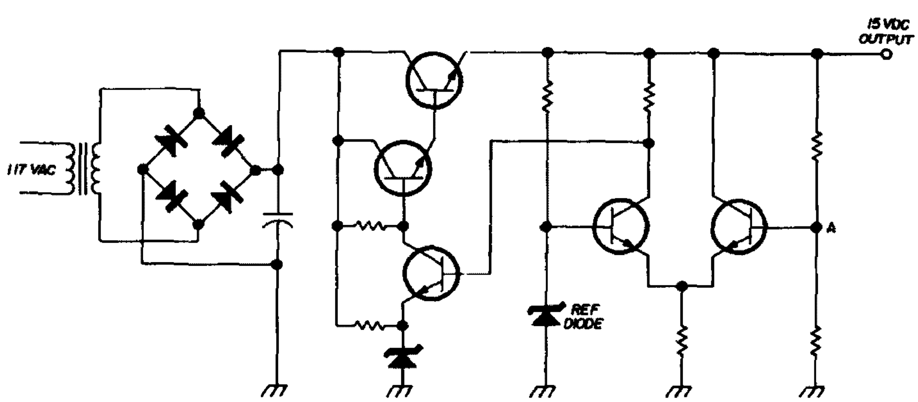
If the modulator and modulation transformer of **fig. 1** are to be eliminated, the output of the rf amplifier power-supply regulator must be made to vary from zero to 30 volts (for 100-percent modulation) in accordance with the audio modulation voltage waveform. **Fig. 3** is a schematic of the power supply shown in **fig. 2**, but the circuit has been modified so that the output may be *varied from zero to* 30 volts by adjusting potentiometer Rp. This, too, is a conventional design.

Removing Rp from this circuit and replacing it with a variable current generator, so that the current through resistor Rf is made to vary with an audio modulating signal, will cause the regulator output to vary with the audio modulating signal. Thus the regulator will be modulated.

## experimental model

**Fig. 4** is a schematic of the modulated supply that was built and tested. Laboratory power supplies provided the input voltages to the regulator. A sine-wave audio generator was substituted for the microphone. The regulator output reproduced the sine wave without notice-

**fig. 2. Regulated power supply that could be used to power the rf amplifier in fig. 1.**



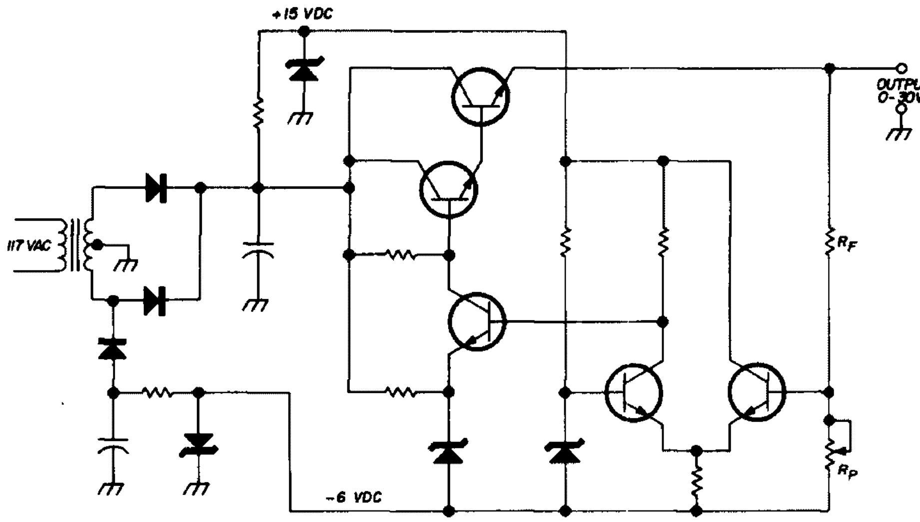



fig. 3. Modification of the regulated supply in fig. 2. Rp is used to adjust tha output voltage.

able distortion with a peak-to-peak amplitude of 30 volts, both under no load and when driving a 60-ohm resistor. Frequency response was approximately 30 Hz to 25 kHz. The microphone preamplifier gain was sufficient to modulate the regulator to 30 volts peak-to-peak by speaking into the microphone in a normal voice. A block diagram of the modified a-m transmitter is shown in **fig. 5.**

Don Jackson, W5QAO, built a two-meter, 15-watt solid-state transmitter using the basic circuit of **fig. 4.** It was necessary to feed the modulated regulator output to both the output stage and the preceding driver to obtain the proper modulation envelope. No other difficulties were encountered, and the transmitter is a compact, trouble-free unit.

Some power dissipation must be sacrificed in the regulator to make the system workable. Since the regulator must be capable of 30 volts peak output, its unregulated dc input must be about 35 volts

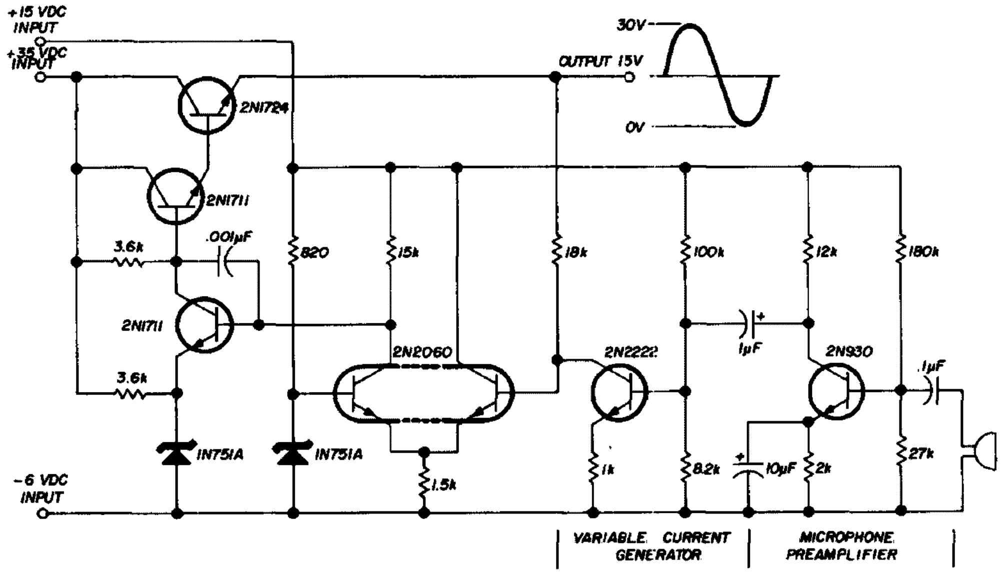



fig. 4. Experimental modulated power supply. The variable current generator is substituted for potentiometer Rp of fig. 1; regulator output then varies with the audio signal.

at an average current of 1 ampere. This is 35 watts input to the regulator, with only 15 watts average power out. The efficiency is about 43 percent.

In the conventional system of **fig. 1**, the combined power out of both modula-

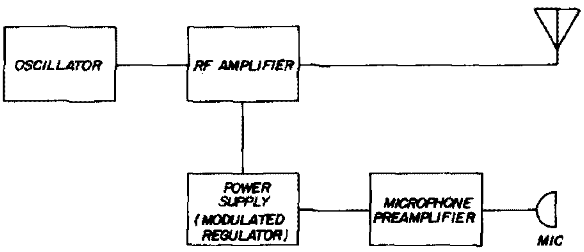



**fig. 5. Amplitude-modulated transmitter with the modulated power supply.**

tor power supply and rf amplifier power supply is 24.6 watts. Assuming an efficiency of 75 percent for these fixed-voltage power supplies, the input power to the two regulators is about 33 watts. Although the efficiency of the modulated regulator appears to be low, the total power input to the conventional system is only 6 percent less than the total power input to the modified system using the modulated regulator.

## applications

Obviously this modulation technique can be applied to any transmitter. In general, the higher the transmitter power, the more desirable this technique becomes, since the eliminated components are more costly as power increases. It's doubtful if this system would be desirable in low-power, battery-operated transmitters of the walkie-talkie type since twice the normal battery voltage would be required. Voltage regulators are not normally used in these transmitters, so a modulated regulator may result in a net increase in components due to the modification, and savings from eliminating a 100-milliwatt modulation transformer may not be substantial.

**ham radio**

# antenna systems for 80 and 40 meters

Some interesting ideas for efficient broadband antennas for the lower-frequency amateur bands

E. H. Conklin, K6KA, Box 1, La Canada, California 91011

**You often hear references** to antenna resonance, but with a lack of indication that the resonant frequency ever was adjusted or measured. Perhaps it is a good thing, for it might lead to confused thinking about the problem.

## resistive mismatch

Take a very simple example—a half-wavelength antenna fed directly with 50-ohm RG-213/U coaxial cable (or 52-ohm RG-8A). The antenna is presumed to present a resistive load of about 72 ohms to the line or $R_R = 72 + j0$. The 50-ohm line needs a load of $R_0 = 50 + j0$, but it is not getting it. The result, 72/50, is a standing-wave ratio of 1.44:1.

The amateur, it is presumed, doesn't actually know that his antenna is at resonance, and doesn't like the resulting swr, so he proceeds to run a curve of changing swr with frequency. Somewhere, at a nearby frequency, the swr bottoms out to a minimum that satisfies him, so he trims the antenna length to put this minimum swr at the desired frequency. Then, he claims that the system is "resonant." Why?

Had the line been a 72-ohm coaxial cable, the antenna load resistance $R_R = 72 + j0$ would equal the characteristic impedance of the line, $Z_0$, and everybody would be happy. The original an-

tenna would still be resonant, and the line would be "flat." That is, regardless of points along the line, or its length, the swr would be 1:1, it would load the transmitter well, and putting in a little more coaxial cable would not change anything.

The original antenna length, fed with the 50-ohm coaxial cable, would worry this chap because of the 1.44:1 swr. He might even add coaxial cable to improve the loading at some new cable length. What he really needs is something to transform 72 ohms at the middle of the antenna to the 50-ohm characteristic impedance of the line. With the transformation the swr at the sending end of the line would be 1:1, and again it would load well. This transformation could have been done in one of several ways, including a transforming bridge balun[1], tapping the line slightly up on a grounded (or ground-plane) vertical, with a stub, with an LC matching network, or with a quarter-wave transmission-line transformer.

## finding resonance

How do you determine the resonant frequency of an antenna if the swr dip does not give it? One quad manufacturer says to insert a one-turn coil in the antenna and grid-dip it; but this may move the resonant frequency down a hundred kHz or more. The *grid-dip frequency must* be checked accurately, such as with a calibrated receiver. And it must be done so the grid-dipper doesn't come too close to the antenna, because that can lower the frequency too.

There are other ways. One is to feed the antenna when it is a continuous dipole, or a grounded vertical, and determine that the currents on both sides of a feed line are identical. A way to do this is to put two rf ammeters in a loop of wire, couple it inductively to an rf source to check meter calibration, and jumper this loop of wire, stretched out with the ammeters in the middle, across part of the antenna with the feedline connected between the meters (**fig. 1**). The frequency that produces identical current in both meters is the resonant frequency.

This method works when the antenna is folded in the middle to produce a tuning or matching stub, as shown in **fig. 2**. However, it is also satisfactory to make a pick-up loop of several small turns of insulated wire, with ends about a foot long, and jumper this across two feet of the antenna. Then a grid-dip oscillator can be coupled to the small coil without it being in series with the antenna. The result is much more accurate. It works on quads, too.

## reactive loads

You may have a resonant antenna with a load impedance (resistive) that is not equal to the characteristic impedance of the feedline. This results in standing waves although the line will present a resistive load to the transmitter when it is any multiple of an electrical half wavelength long. This load will be that of the antenna, which may or may not make the transmitter happy.

At any intervening length of feedline the actual impedance will be complex—that is, it will have a resistive and a reactive component. Some lengths of line may produce a combination of resistance and reactance that the transmitter cannot load because of matching network limitations. In this case, you can live with the high swr by adding sections of coaxial line until a length is reached that provides a more satisfactory load to the transmitter. The actual swr, however, remains the same.

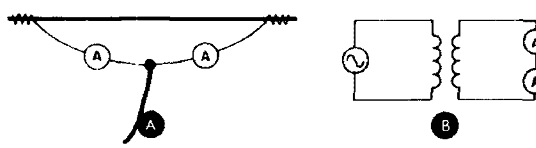



**fig. 1. Determining resonance with rf ammeters. When the current through each ammeter in A is the same, the antenna is resonant. In B, the ammeters are placed in series to check that both instruments read the same.**

Another solution is to add a series or parallel capacitor or inductor to the line to cancel the reactance.

When my Henry 2K is used on the wrong end of the 80-meter band, the tuning and loading controls no longer are where they were when feeding a dummy load; the tuning control is at maximum, and the loading control alone is adjusted for a plate-current dip, allowing no control over loading. As coaxial cable is added to the line the loading improves and the controls on the amplifier move toward the setting when feeding the dummy load.

## bandwidth

If a dipole is used over a 12 to 14 percent bandwidth in the 80-meter amateur band there are problems due to the change in complex antenna impedance presented to the feedline as the frequency is moved from one end of the band to the other. When the antenna itself is mistuned in order to terminate the line with a suitable amount of reactance to minimize the swr it may operate at a point on its reactance curve where small changes in frequency create large changes in reactance. This results in a narrowing of bandwidth, which might be defined as the bandwidth for an acceptable swr. A result of this condition was the development of in-band rf traps,[2] the use of coil switching at the antenna,[3] end clip-on wires, and other means of enabling the antenna to be used over the entire 80-meter band. One approach[4] used series capacitors, a stub of fanned-out wires to create a real 10 percent bandwidth or more with a 1.5:1 swr.

In the case of the inverted-V antenna, fortuitous height and apex angle may cause the antenna to present a 50-ohm load at resonance.[5] Some increase in bandwidth may result as compared with a horizontal dipole in the absence of matching or broad-banding. Broadband impedance-transforming bridge baluns have been developed[1] for use when the antenna impedance at resonance does not match the transmission line properly. However, solutions by step tuning or by true broad-banding continue to be required, particularly for those whose antenna facilities are limited. Many of us need an antenna that covers 40 meters and both ends of 80 meters. We now have simple solutions for the 40-meter band and for spot frequencies of 3525 and 3825 kHz more or less, using the in-band traps or dual 80-meter wires, but we must develop a more general approach.

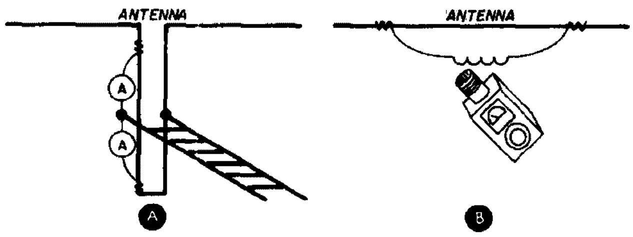



**fig. 2. Determining antenna resonance with series rf ammeters in a matching stub, A, and with a grid-dip oscillator, B.**

With the current 5-band DXCC interest in 80 meters, better solutions are needed. Frequently, one is caught with the end loading wires clipped on or off, presenting a 15-to-1 swr or worse at the necessary operating frequency. The system can be loaded with a suitable matchbox, or addition of coaxial cable, or lumped reactance, but there are added losses with such a high swr.

## stub matching

Before leaving the general comments, let's take a quick look at some of the simple theory of stub matching. If an antenna is resonant and presents a satisfactory resistive load equal to the characteristic im-

pedance of the line, there is no problem. However, if the length of the antenna (or frequency) is changed it becomes reactive. If this length is selected to bring the resistive component of the complex impedance of the antenna, $Z_R = R + jX$ to a point where the R is equal to the characteristic impedance of the transmission line, $Z_0$, a reactance (equal and opposite to the antenna's reactance) placed across the line will result in a match. The necessary reactance can be a stub. Usually a shorted stub is preferred. If you know the $R + jX$ at the center of the antenna,[6] the stub length that can provide the cancelling $-jX$, can be calculated, read from a graph, or obtained from the Smith chart.

Let's look at the situation in another light; **fig. 3** shows a quarter-wavelength stub with a line attached at a point that gives a proper termination to the transmission line. The impedance of the shorted stub in one direction from the transmission line tap, a, is equal to that

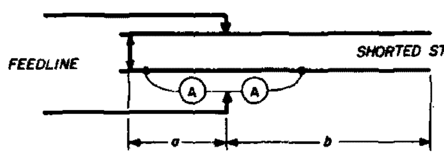



**fig. 3. Line matched to stub; the reactance of the shorted section (A) at the line taps equals the reactance of the open section (B).**

of the open stub in the other direction, **b.** The current flowing in one wire of the stub will be equal just below the line tap, and just above the line tap, when the stub is resonant. The open part of the stub can be bent out to form a dipole antenna, with the feedline conveniently located at the bend. This form of stub may shorten the dipole.

On the other hand, if you have an arbitrary length dipole and wish to feed it you may back down the line to the one or two points, a in **fig. 4,** in the first half-wavelength toward the transmitter which has a resistive component (in $R + jX$) equal to the characteristic impedance of the line. At that point, attach a stub, **b,** which has an equal and opposite $-jX$ reactance, thus cancelling the reactance and matching the line with the resistive load from that point back to the transmitter.

## smith chart

The Smith chart[7] is a useful tool. A circle on it, centered on the center of the chart, is an swr circle for a lossless line. Points around this circle, which covers a half wavelength, gave the resistive and re-

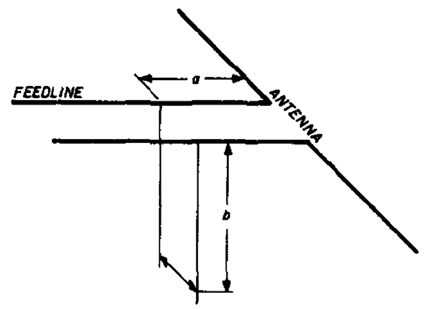



**fig. 4. Method of stub-matching a reactive antenna. All lines can be coaxial cable.**

active components of the complex impedance along the transmission line.

In reverse, any value $R + jX$ complex impedance can be plotted on the chart, such as the changing resistance and reactance as some circuit is fed at different frequencies. When these impedance points are connected together with a line it will be seen what range of frequencies falls within an acceptable swr. The Smith chart also facilitates adding reactances in order to bring some part—or a greater part—of the connected points within the acceptable swr circle, by moving a plotted point along a resistance curve by the amount of the added reactance.[8]

So far we have dealt with resistance, impedance and reactance. Sometimes problems are more readily solved in the reciprocals (these values divided into 1)

called conductance, admittance and susceptance. Whichever form is used, it is sometimes convenient to divide $R + jX$ by the characteristic impedance of the line, use the result in chart work or calculations, and then multiply again by the line's $Z_0$ to obtain the actual values of $R + jX$. This process is called "normalizing." One advantage of normalizing is that the value applies equally to lines of different $Z_0$ without replotting. Smith charts are available with a normalized 1 at the center for normalized data, and others are available with 50 at the center for direct use with 50-ohm systems.

about four turns were added; for 80-meter cw, about 9 turns were added.

An swr bridge showed something like 1.5:1 but there was no transmission line to have a standing wave. This actually was a measurement between the normal 52-ohm termination in the swr bridge, and the actual resistive load presented by the antenna. At any rate, the Heathkit SB200 and Henry 2K both fed the antenna easily, which was only 15 feet above ground at its midpoint. Nevertheless, cw and ssb contest contacts were made into far South America, Malaysia, Singapore, and the like on 80 meters.

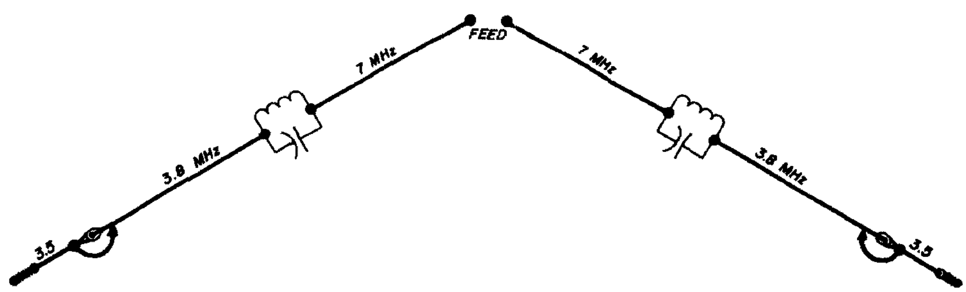



**fig. 5. Trapped inverted-V with clip-on ends. Coaxial connectors ara more reliable than alligator clips.**

## simple two-band antenna

When I was becoming more active in amateur radio after retirement, I didn't appreciate that the problem of coax-fed antennas for the lower bands might be difficult. With some memory of a chap named Marconi, and disregard for the 200-ohm resistance between rods driven in the ground ten feet apart, I put up an L antenna which was resonant at 5/4 wavelengths to the 40-meter trap, and at 3/4 wavelengths on 80 meters to the far insulator. The antenna was led to the transmitter output terminal with **no** coaxial cable at all, and **no** matching or coupling device. The wire was trimmed for the high end of both bands. Then, to move it to the low end of both bands, about ten turns of two-inch Air-Dux coil were mounted on the wall, and provided with a shorting clip. For 40-meter cw,

The main problem was the necessity to put 16 bypass capacitors and one rf choke in the electronic keyer, and to use RG-58/U cable instead of shielded audio cable to the exciter key jack to prevent rf from interfering with the keying. The antenna system was so simple, really, that it was a pity that use of the land was prevented by a building program. With the local decomposed granite soil, and pavement over almost all of the ground, all proposals that I use a vertical antenna and a radical ground were quickly rejected without a fair test.

## two-band dipole types

The common two-band inverted-V or dipole is the trapped wire, **fig. 5.** This antenna may have considerable interlocking tuning between the sections. Furthermore, *if any dead-end wire is twisted back*

around itself, thus providing more wire if needed, it may load that section of the antenna unless it is shorted to the active wire by soldering or clamping with a *Kearny clip. Even a loose fold-back may* change the tuning by acting like a "fat" wire.

Another version is the use of two wires, each being a dipole for one of the bands; these wires can be at different angles, or spaced by dowel-rod spreaders, using a single support at each end, as in **fig. 6.**

These antennas may provide adequate coverage of the 40-meter band, but the tend to shorten the resonant length by as much as 20 percent.

Everett[9] goes through the design of a fat antenna with a bandwidth of about 30 percent *within* an *swr of 1.25:1. The* length/diameter ratio of the conductor was selected for a resistive component that matches the transmission line; then a shorted stub was connected across the line at the antenna to cancel out the reactance.

John Kraus, W8JK,[12] gives the resistance of fat antennas which runs in the vicinity of 80 ohms at the center of a half

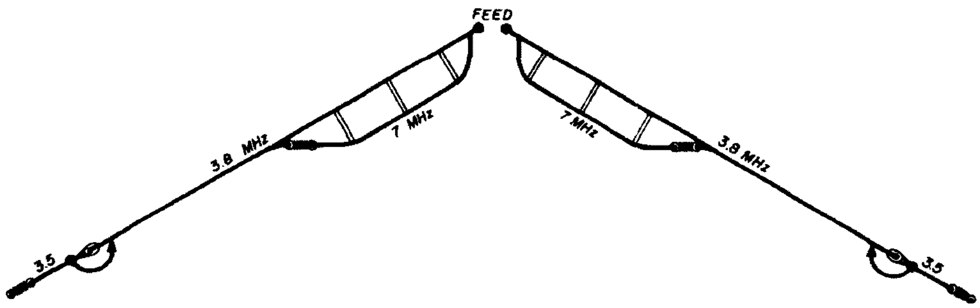



**fig. 6. Two-wire, two-band inverted-V with clip-on ends.**

width of the 80-meter band may be too much for them unless there is excessive loss in the line or balun. Line losses can be compared with those existing during earlier measurements, by logging the swr when the far end of the line is shorted.[10]

## broad-banding

On one band only, there are several ways to broaden the antenna. One way is to make the wire fat, in effect, compared with the length. Several spaced wires can be used, including the old "cage."[11] A remarkably effective and convenient form is to use several wires, held apart half way between the center insulator and the end by some type of light-weight spreader, forming a diamond in each quarter-wavelength. Still another way is to fan out two wires, or three, supporting the far ends from different points. All these wavelength, to 200 to thousands of ohms for the center of a full wavelength. These lengths are much shortened due to the shape. The convenient impedance explains why many very broadband fat antennas actually are vertical half-waves or horizontal full-waves.

Coleman[4] has investigated the use of parallel or series reactances in the form of capacitors, inductances or line segments, to *obtain more* bandwidth in a particular antenna. In connection with parallel-resonant stubs (usually slightly longer than a quarter wavelength) placed across the line at the antenna, he says: "The broad-banding property seems not to have been exploited. A resonant antenna has a negative susceptance (1/X) slope with respect to frequency, which can be cancelled with a properly chosen circuit over a considerable range of fre-

quencies. The most favorable antenna curve is one which has a resonant conductive component (1/R) just less than 2. The stub portion of a bazooka (quarter-wavelength-line type of balun) is ideally suited for bazooka matching and balancing of a balanced resonant circuit to an unbalanced line." Linear baluns thus formed of coaxial or triaxial[13] cable may be very useful in broad-banding an antenna without added hardware.

Meier[4] has put together the work of Coleman on the use of capacitors or inductors, and stubs, to obtain broad-banding. With simple antennas, and using a two-element broadening arrangement consisting of a series capacitor and a parallel stub, he obtained bandwidths far greater than our 80-meter band. Then, he explored the use of fan elements (**fig. 7**) which are physically short, and found that it was necessary to use only the parallel stub in order to obtain adequate broad-banding. Three wires, connected together at their far ends, were satisfactory; the length of the center wire plus half the length of the wire connecting the ends, was approximately a quarter wavelength.

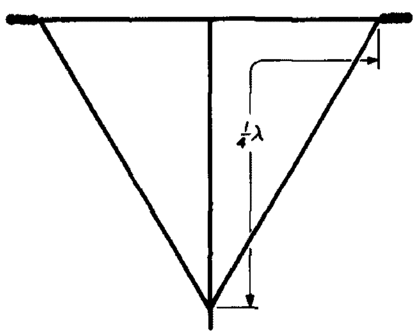



**fig. 7. Fan monopole, physically short as shown. Two can be used as a dipole.**

The unusual bandwidth resulted partly from the complete loop that the antenna impedance makes on the Smith chart. The spiral locus is also typical of a series line transformer with a length that exceeds a quarter wavelength.[9]

It is customary practice in broadband designs to sacrifice a perfect match at the midfrequency in order to gain bandwidth within the acceptable swr. For a fanned-out antenna and linear balun giving an swr less than 2.5:1 over the 80-meter band, (see the Radio Handbook, 17th Edition).

## double-humping

Somewhat like broad-banding is the design of an antenna that will produce an acceptable swr at two different frequencies within the 80-meter band. One way is to clip additional wire on the ends of the antenna, but this sometimes proves to be inconvenient. A second way is to use in-band traps to do the clipping-on automatically.[2] It may be possible to use line sections as automatic switches[14] even if a coiled coaxial line is used on only one side of the center of the antenna to perform the switching function, such as shorting a series capacitor or changing the length of a stub to its shorting bar or shorting line. I have tried separate dipoles cut for two frequencies in the 80-meter band (**fig. 8**). This worked satisfactorily except for the amount of wire in the air.

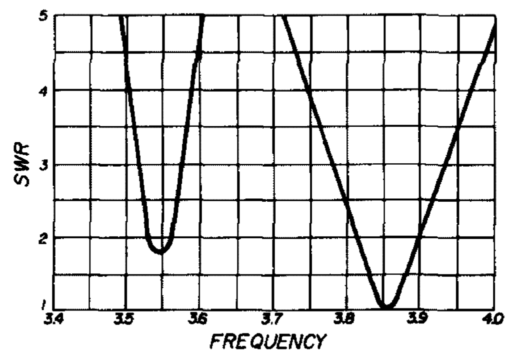



**fig. 8. 80-meter antenna using two dipole wires within the band producing a double-humped swr curve.**

Two tests were made to solve the problem of midband operation. With the clip-on ends, leaving one on and one off, put the antenna swr dip in midband. This has not yet been tried with the in-band traps,

but clip-on wires hanging from the inside end of the traps successfully moved the position of the swr dip. This idea was first considered as a means of lowering the frequency of the ten- or fifteen-meter sections of the driven element of a Telrex Yagi without affecting lower bands appreciably, and without adjusting the aluminum tubing.

## two bands broad-banded

Many of us have the more difficult problem of covering the 40-meter band at both ends, and the 80-meter band on both cw and phone. This tends to make the problem more difficult. The trapped 40/80 inverted-V with clip-on ends for the 80-meter cw frequencies which has been in use at K6KA for several years is somewhat difficult to adjust initially because of the loading effects of the 80-meter sections upon the 40-meter section inside of the 40-meter Hy-Gain traps.

The need for clipping on wires was found unnecessary when an 80-meter phone dipole was placed in parallel with the old 80-meter cw dipole which was trapped for 40 meters. However, there continued to be some interlocking adjustments.

An alternate plan used one dipole from the center of the antenna to the 40-meter trap, and then two fanned wires from the trap cut to different lengths for the ends of the 80-meter band. This caused a remarkable shortening of the wire length in addition to that caused by the inductive reactance of the trap, and only one swr dip was located within the transmitter's frequency range. The narrow frequency range of current exciters makes it difficult to locate the swr dips when they fall well outside a band.

Using two dipoles spaced six inches by dowel rods, the 40-meter wire below the 80-meter wire, and in-band traps near the end of the 80-meter wire, W6JKR found no interaction and fully satisfactory operation.[2]

Obviously, the in-band traps can be eliminated, too, by using three wires: one for 40 meters, one for 80 cw, and one for 80 phone. Such an arrangement was put up at WB6ITO and WA7NAR using a wooden mast and a four-wire cage dipole on X-shaped spreaders. The wires are not connected at the far ends, but the lengths are cut for different frequencies in order to produce a fairly flat swr over the entire 80-meter band. One wire presumably can be cut for 40 meters. The problem of adjustment might be simplified by clipping extension wires on the three (or two) 80-meter dipoles not being adjusted, in order to move them out of the range of the dipole being adjusted. Actually,

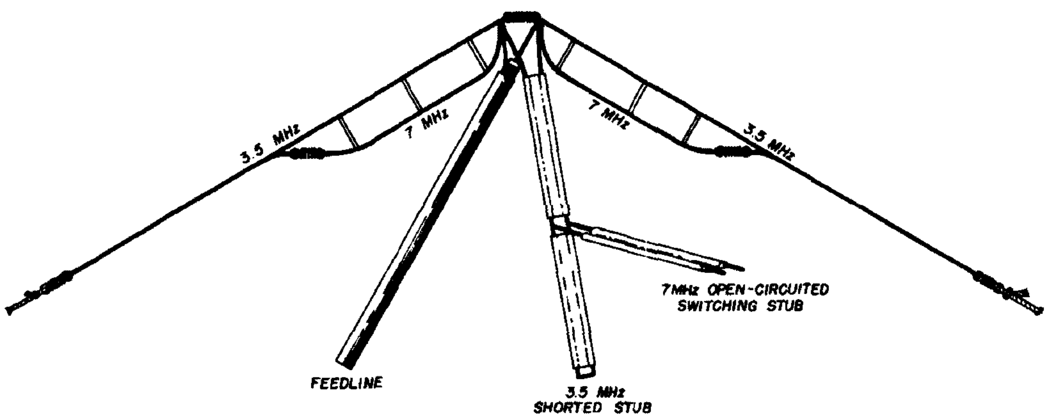



**fig. 9. Broadband 80- and 40-meter inverted-V with tv shielded foam line shorted stub, and quarter-wave 7-MHz switching stub. This can be part of a bazooka balun.**

the adjustment is not to "resonance," but to a low swr at the dipole's assigned frequency in the band.

## general solution

So far several satisfactory one-coaxial-cable two-band antennas have been mentioned, including those with a separate wire for 40 meters, and two-point or wide-band coverage of 80 meters. If two coaxial cables are considered, the problem of obtaining broadband coverage of 80 meters is not difficult if a quarter-wave stub or bazooka balun is used for broadening.

For full coverage with a single feedline, other than the two cage dipoles described by Covington,[11] or a single cage which might permit trapping one wire for 40 meters, it appears that the most satisfactory design will use a combination of methods such as those discussed above and will be useful throughout both 40 and 80 meters with an acceptable swr. The most promising possibilities appear to be:

**1.** A 40-meter wire and an 80-meter wire, broadbanded by a parallel quarter-wavelength (or slightly longer) stub or bazooka whose length is automatically switched between bands by applying a quarter-wavelength open-circuited line at the position that will shorten the stub to the 40-meter length (**fig. 9**).

**2.** A fan dipole with 40-meter traps in all of the wires, and a parallel quarter-wavelength stub (or slightly longer, possibly made of shielded tv foam cable coiled up) or bazooka with the automatic switching discussed above (see **fig. 10**).

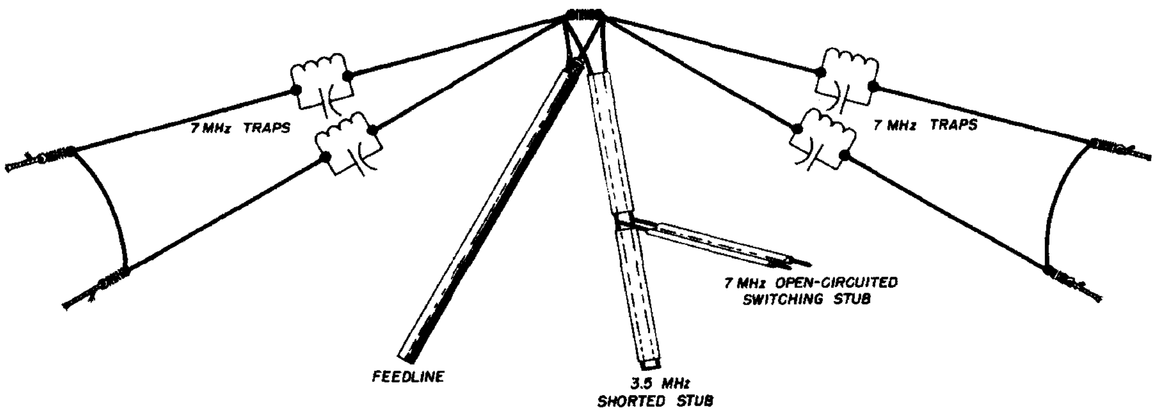



**fig. 10. Broadband 80- and 40-meter fanned inverted-V, trapped for 40 meters. Stubs perform the same as in fig. 9.**

### references

1. W. I. Orr, W. Sayer, "Wide-band Bridge Baluns," *ham radio*, December, 1968, p. 28.
2. R. Polansky, "Low-band Converted-vee Antenna," *ham radio*, December, 1969, p. 18.
3. N. E. Handel, "Novel Antenna for 80 and 40 Meters," *QST*, February, 1969, p. 40.
4. Bennett, Coleman and Meier, "The Design of Broad-band Aircraft Antenna Systems," *Proceedings IRE*, October, 1945, p. 671.
5. D. W. Covington, "Radiation Resistance of Inverted-V Antennas," *QST*, October, 1968, p. 36.
6. Strandlund, "Amateur Measurement of R + jX," *QST*, June, 1965, p. 24.
7. R. Hall, "Smith Chart Calculations for the Radio Amateur," *QST*, January, 1966, p. 22; February, 1966, p. 30; March, 1966, p. 76; June, 1966, p. 40.
8. K. Amis, "Antenna Impedance Matching," *CQ*, December, 1963, p. 65.
9. Everitt and Anner, "Communication Engineering," McGraw Hill (1956).
10. Cholewski, "Some Amateur Applications of the Smith Chart," *QST*, January, 1960.
11. D. W. Covington, "Inverted-V Radiation Patterns," *QST*, May, 1965, p. 81.
12. J. Kraus, W8JK, "Antennas," McGraw-Hill, p. 245.
13. J. Schultz, "Cable Pickup and Shielding," *73*, September, 1969, p. 36.
14. "Radio Communication Handbook, 4th Edition," Radio Society of Great Britain, 1968, p. 1373.

**ham radio**

# quick band change

## from six to two meters

If you like to operate on both six and two meters, here's an equipment changeover scheme that is practically instantaneous

W. G. Eslick, KØVQY, 2607 E. 13th Street, Wichita, Kansas 67214

My problem was how to get back and forth between 6 and 2 meters with this rather complex array of equipment: An Ameco TX62 (6- and 2-meter output on one coax connector); two converters, indicating lights; on-the-air light; and a speaker switch. One way would be to change a lot of coax cables, power cables, converter connections, etc. Meanwhile, the skip would have come and gone while thrashing around trying to get set up on the other band. A much quicker way is shown in **fig. 1.** Here's how I did it.

I had three surplus coax relays that came from an old vhf ARC-5 and a 12-volt dpdt relay. Coax relay 4 (**fig. 1**) was ordered from K8ZES, whose ad appeared in the August, 1955 issue of the *Vhfer.* These were two dollars post-paid at the time. They're smaller than standard coax relays. I used this to change my receiver input to either converter. Although these relays are 24- or 28-volt types, they've been working fine on 12 volts for three years. If you have a 28-volt coax relay, try it on 12 volts before you spend your hard-earned cash on another.

With all coax relays unenergized, the station is on 6 meters. Switch 2 is mounted on the mike. It energizes the on-the-air light (red jewel) and also energizes the dpdt relay, which disconnects the speaker and shorts the audio through a 10-ohm resistor. The other relay contacts ground the cathodes in the TX62. Coax relay 2 is

**table 1. Parts data for the 6- and 2-meter switching circuits.**

| | |
|---|---|
| T1 | Transformer, 110 V primary; 18 V 2½ A secondary |
| D1 | Stud-type (unmarked) rectifier from Polypaks |
| Bulbs | 12 V bayonet type number 1820 mounted in holder with colored jewel |
| Coax relays 1 through 3 | war surplus salvage |
| 4 | see text |
| SW1 | dpdt toggle switch |
| SW2 | modified microswitch mounted on microphone |

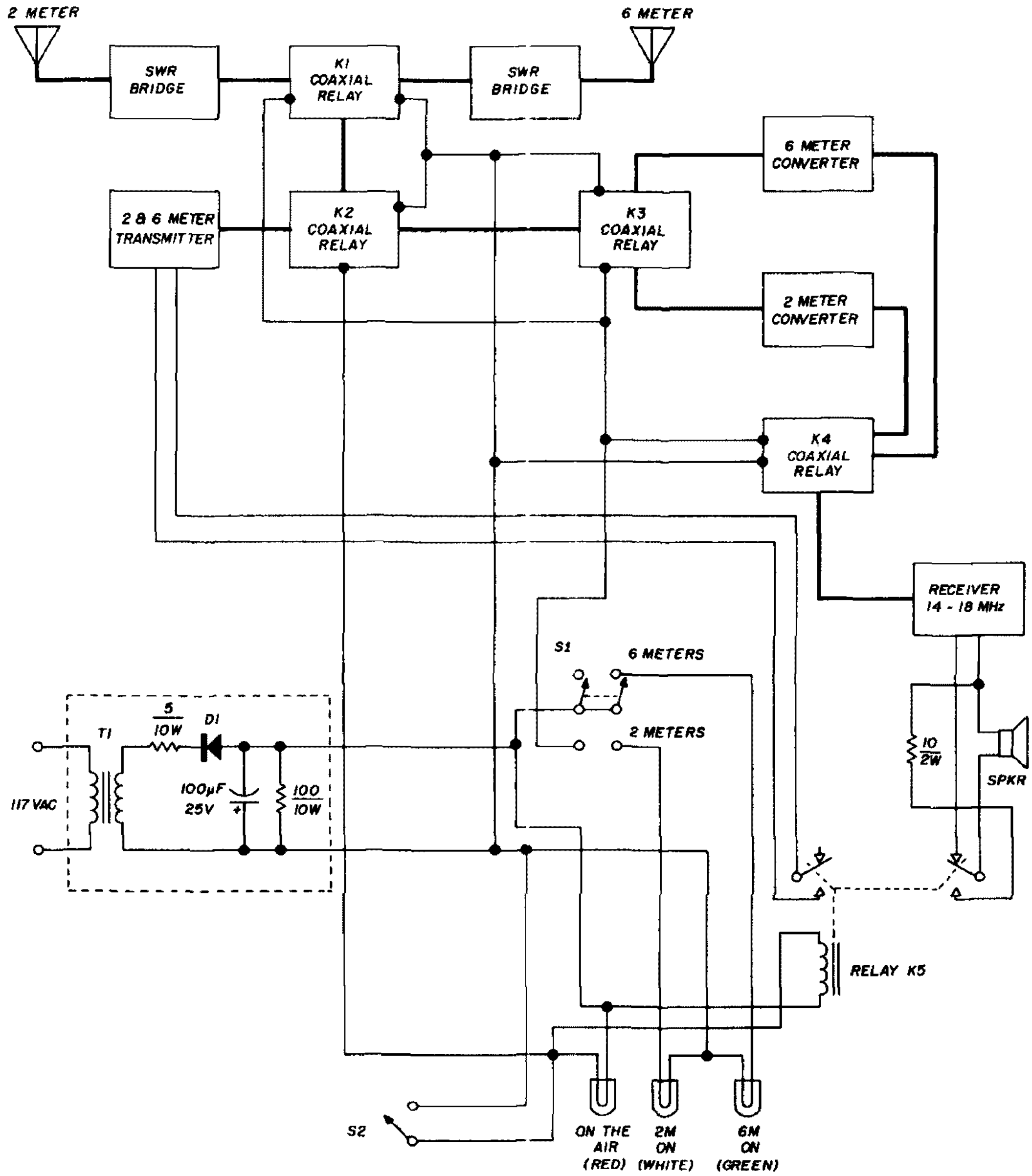



**fig. 1. Switching circuit for quickly changing between 6 and 2 meters. Coaxial relays are coaxial dc-operated types. Heavy lines represent coaxial cables.**

changed from receive to transmit through the coil of relay 5.

Throwing toggle switch 1 to the 2-meter position lights the 2-meter indicator light (white jewel), selects the 2-meter antenna via coax relay 1, connects the antenna circuit to the 2-meter converter via coax relay 3, and connects the output of the 2-meter converter to the receiver. With toggle switch 1 in the 6-meter position, everything is connected for 6 meters, and the 6-meter light is on (green jewel).

This may sound complicated to some and may be too much of a lashup for others, but it's easy to work. With switch 1 up and the green light on, after I throw the master switch that turns everything on, I'm listening on 6. I press the mike switch, and I'm on the air. If I want to go to 2 meters in a hurry, all I have to do is place toggle switch 1 in the down position, the white light goes on, and I'm on 2 — just that simple and fast.

Additional information on the parts I used is given in **table 1.** (Other parts are as marked in **fig. 1.**) If you're interested, the swr bridges were made from an article entitled "Vhf Monimatch" in the **Radio Amateur's Vhf Manual.**

**ham radio**

# the ham notebook

## converting the hallicrafters SR-160 to an SR-500

While considering the purchase of either a used Hallicrafters SR-160 or an SR-500, I was looking over the two instruction manuals. The schematics showed that both units were identical with the exception of the final amplifier tubes. The circuits were the same, and even the same components were used in both final amplifiers. Because of this similarity, it's a simple job to convert the SR-150 to an SR-500.

Before converting the transceiver, you should think about the new power requirements. It's not economically feasible to convert the SR-160 power supply, since this would require a new transformer, new diodes and new liter capacitors for the high-voltage supply. The most economical way is to sell the old SR-160 power supply and buy a used SR-500 power supply. According to current used-equipment prices, it should be possible to swing this deal with a cash difference of about $25.

The other approach is to build a supply for the converted SR-160/500 from scratch, and you could keep the SR-160 supply for later conversion to original configuration for resale.

If you build your own power supply it must provide the following voltages:

280 Vdc, 100 mA
750 Vdc, 50 mZ
-80 to -130 Vdc, 10 mA
12.6 Vac, 5A

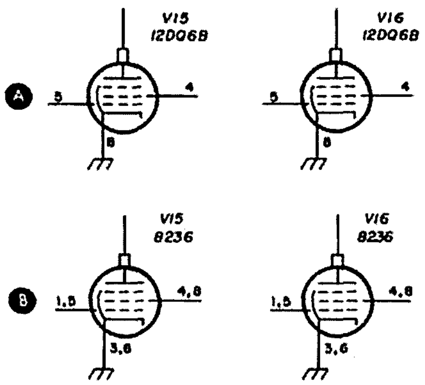



**fig. 1. Converting the Hallicrafters SR-160 to an SR-500. Original SR-160 final amplifier wiring, A, and the SR-500 circuit.**

Once the power supply is taken care of, the modification of the SR-160 is as follows.

Remove the two 12DQ6B final amplifier tubes from their sockets. Rewire the socket connections to the final amplifier tubes as shown in fig. 1. Install two 8236 tubes in the final amplifier sockets.

Readjust the bias for an idling current of 100 mA, using the method outlined in the SR-160 instruction manual.

Due to the similarities of the two equipments, the SR-160 manual and the SR-500 manual are identical. Note the changes made in the final amplifier socket wiring and the new idling current required in the bias adjustment.

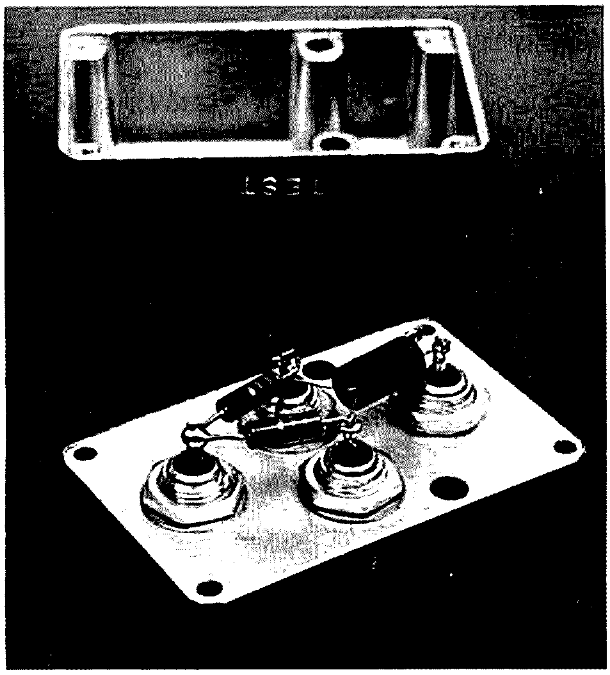
As a matter of interest, I noted that the picture of the SR-160 bottom chassis view (Figure 13 of the manual) had been used in the SR-500 manual. It was possible to make this identity by checking the pin connections to the final amplifier tube sockets.

If you have an SR-160, this is an easy way to beef up its output a little.

**Al Brogdon, K3KMO**

# impedance bridge

Here is a useful little bridge that can be used to compare the impedance of an unknown antenna (or network) to a known termination over the range from 2 MHz to 900 MHz. By changing the value of the



Construction of the impedance bridge. Unit must be housed in a metal box.

termination, impedance can be checked in the range from 25 to 500 ohms; to check the match of a 50-ohm antenna for example, a 5-ohm load would be placed across the *termination* terminals. When the impedance across the *test* terminals is equal to the

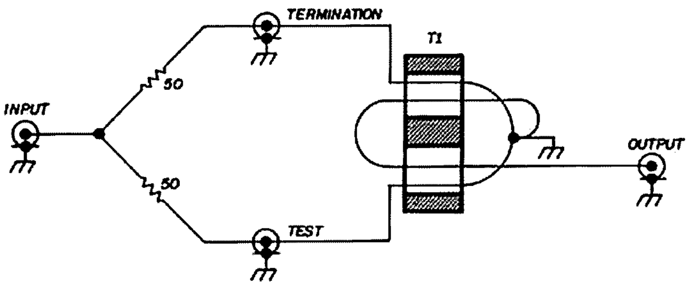



fig. 2. Impedance bridge compares "termination" impedance to the "test" impedance. Transformer T1 has two windings; each winding is one turn through a small double-hole toroid, the windings connected as above.

impedance across the *termination* terminals, there is no output.

This bridge may be used with a sweep generator and oscilloscope to display a circuit's resistance characteristic over the desired bandpass. To determine the value of an unknown impedance connected at the *test* socket, a calibrated variable resistor may be used at the termination terminals. When building this simple instrument, make sure all lead lengths are equal. It is not necessary for the impedance of the measuring equipment connected across the *input* and output terminals to have the same value as the unit under test.

**Oliver W. Swan, W6KZK**

# instant replay

Does your tape recorder sit idle while you're on the air? All you need is a continuous tape loop, of as long a length as practical in your shack, made by slicing and joining both ends of a length of tape. With appropriate "jury rigging", such a loop can run forever without snagging. Continuous-loop automotive cartridge recorders such as the Sony TC-8 could also be used.

Whenever you do a bit of listening on the bands, run the recorder, either directly from the receiver's speaker output, or simply through the recorder's mike, placed near the speaker. Leave the machine in the "record" position; you will, at any given moment, have the most recent sounds (or minutes) of received signal, depending on loop length, which can be instantly replayed by quickly switching to the play mode. Why? Imagine this:

**1.** As a novice, you're trying to copy a signal slightly above your maximum cw receiving speed, to help improve your own speed. You got nearly all, but what did you miss? Flip to "play", cut the recorder's speed from maximum where the loop was originally run down to minimum. If your recorder has two speeds, you'll have halved the original wpm to a more comfortable level. If it has three speeds, you can cut down to one-quarter of the original speed.

2. You're a DX hound, on cw or phone, and you think you've just heard an unbelievably rare one. Or was it? Play it back again, maybe twice, before trying to call. Too bad! It wasn't ZK3JO . . . just K3JOZ.

3. You're handling traffic when a noise in the shack distracts you. You've missed an important part of the message. Ask for a complete repeat? That won't be necessary if you've caught the missing section on tape.

Of course, a tape loop isn't absolutely necessary; you could record the entire operating session on a full reel to be replayed when a log entry is to be clarified, and re-recorded on the same tape during the next session. But a loop, especially a short one, eliminates the time and uncertainty of fast rewinding. Experiment with different loop lengths (times) to find the length best suited to your operating habits. Using a loop is like having an insurance policy. You'll begin to appreciate it the first time you really need it.

**Bob Hirschfeld, W6DNS**

# using an outboard receiver with the SB-100

The following idea is offered as an addendum to a similar item that appeared in an earlier issue of **ham radio.*** In my case, it was a simple matter of rewiring two sets of contacts on my Heath SB-100 antenna T-R relay so I could use a separate receiver with the unit. I wasn't concerned with switching the two audio outputs to a single speaker, as shown in the previous article, so no provision was made for this.

The before-and-after circuits are shown in **fig. 3.** I'm not familiar with the differences between the SB-100 and the SB-101, but if the T-R relay setup is similar in the two units, the modification should also work for the SB-101.

**Bill Clements, K4GMR**

*Jim Fisk, "Using an Outboard Receiver with a Transceiver," *ham radio*, September, 1968, p. 12.

# soldering tip

Dig around the lumber yard for a scrap of lumber with a big end-grain knot, the sappier the better. Make a depression in the knot to hold a drop or two of solder. Then, when you want to clean and tin your iron, just rub the tip on the knot. Cover the bottom of the knot to keep it from glopping up your bench.

**James T. Lawyer**

# LPY antenna

Unfortunately two of the driven elements were transposed in **fig. 5,** page 13 of the July issue; the positions of the third and fourth driven elements should be reversed. If the antenna is built as shown there will be a serious loss of gain.

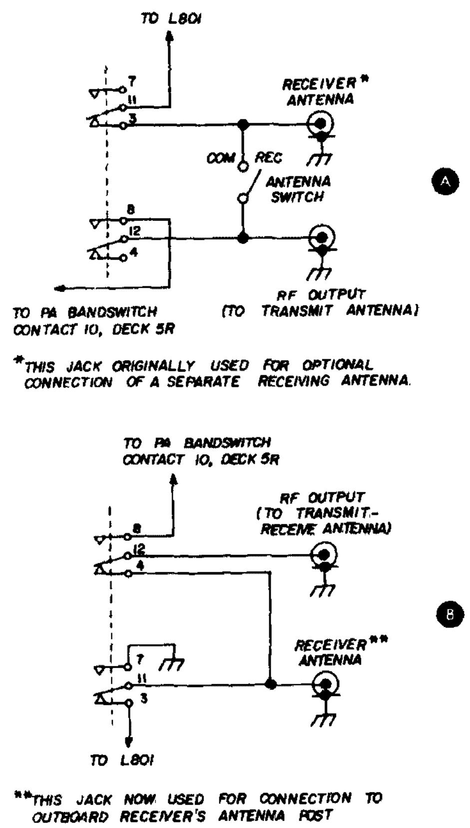



**fig. 3. Original, A, and modified circuit, B, showing SB-100 T-R wiring for outboard receiver; no new parts are required.**

# a mobile mount bracket

Serious thought must be given to mounting a transceiver in the front seat of today's new cars, because the bracket normally accompanying the equipment won't allow suffi-

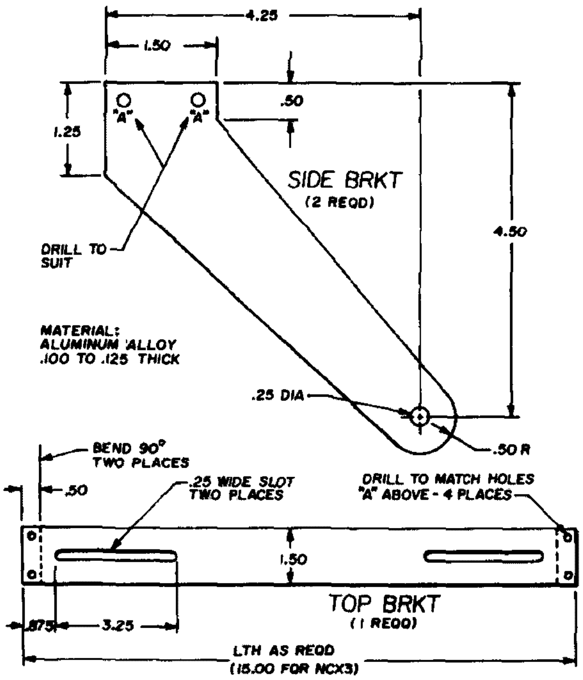



**fig. 4. Homemade mobile mount bracket.**

cient clearance above the hump in the floorboards when the equipment is bolted under the dashboard.

I found it necessary to design a mobile mount bracket with a slightly different shape to fit my NCX3. Also I recalled the many times my wife had caught her stockings on the arms of the old bracket when I had the rig out of the car. So with all that in mind, the design in **fig. 4** was created, and the side arms were made removable.

I constructed the bracket from a 1½-inch piece of 5052-H32 aluminum alloy with a hardness index of 46. However, any hard aluminum that can be bent 90° will suffice. I cut it the same length as the bracket that came with the equipment so that when ½-inch right-angle lips were formed at each end, it fitted over the transceiver case. PEMNUTS with 6-32 threads were fitted into the ½-inch lips, and, of course, the body of the bracket was cut so that when fastened under the dashboard, the bracket could be slid either way for proper positioning. The arms were fashioned as illustrated, but as each car is different, their angle may vary. This can be determined almost by the naked eye. The arms are fastened by 6-32 screws that go through the neck of the arms and into the PEMNUTS.

The result is a very firm installation. When the rig is out of the car, the arms are put into the glove compartment, and the wife's stockings and shins are safe. WA4KDI did the mechanical design.

**Gay E. Milius, Jr., W4NJF**

# pentode replacement

Many amateurs are updating their old equipment by replacing vacuum tubes with semiconductors. Bipolar transistors are used occasionally, but the trend seems to be to junction fets and mosfets, with mosfets taking the lead. Jfets are a little harder to use, but not much, and they are still lower priced than rf-rated mosfets. The circuit shown in

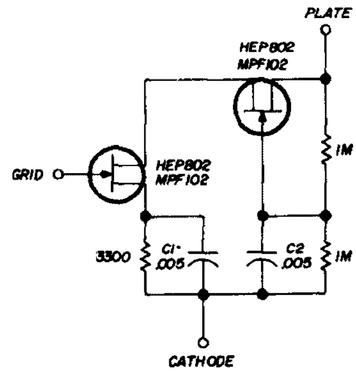
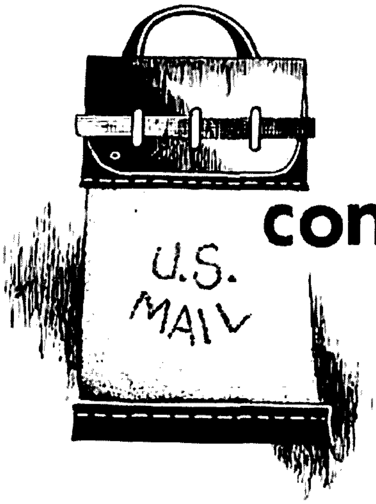



**fig. 5. A simple solid-state circuit that can be used to replace vacuum tubes in simple amplifier circuits.**

**fig. 5** is a simple cascode circuit that may be used as a simple plug-in replacement for most pentode vacuum tubes. The circuit shown here works well up to 30 MHz; for higher frequencies bypass capacitors C1 and C2 should be reduced to 0.001. When using this circuit, make sure that the voltage at the "plate" terminal does not exceed 50 volts. Current drain of the circuit is about 600 microamperes.

**Jim Fisk, W1DTY**

# comments




**Dear Mr. Bonadio:**

I am one of "those Engineers" you are talking about. I would like to make some comments regarding your antenna, and also about science in general. Many discoveries are made by people not trained in the field, and anyone who makes one deserves a lot of credit because he starts with a distinct disadvantage. He has to try everything: things that can pay off as well as those things that cannot because of physical laws against them. An additional disadvantage is that the untrained inventor fails to see the commonalities inherent in many designs that look quite different.

There is no question in my mind that the antennas you describe work as described and in many situations will give better results than the beam antenna next door. That still does not mean that they are superior antennas in general. This, I think, is the reason why you could not sell them to manufacturers.

The "square diagonal" antenna has the characteristics of a dipole. You give directions for switching its (horizontally polarized) pattern by 90 degrees by means of relays. By adding two more positions on the switch you can switch the beam 45 degrees at a time. Just open up one set of dipoles at a time. The antenna

**Mr. Bonadio replies:**

*Yes, I've tried to violate some of what we believe to be physical laws, because I don't trust everybody else's beliefs either. What I was originally trying to do was to broadband a doublet. I did. I then went to logical extremes. Then I could switch it and keep it in tune—at fabulously low Q. Then the ridiculous reports started coming in. These were on comparative strengths from DX and fading comparisons in round tables. This was from two dimensional elements. So, naturally, I tried three dimensions. The ratio of reports improved. Enthusiastically I tore down my old comparison antennas. Then I idealized the three dimensional effect in the six-element space dimension antenna, and found it better than the four- or eight-element systems.*

*I sent the editor of ham radio a list of about 100 ridiculously good reports from my modest operating before he would accept my article. Operated correctly, these antennas are actually very effective. However, sales to a manufacturer are based on that manufacturer needing the item for his profit picture. After amateurs prove that these antennas have a desirable place in sales, manufacturers can be interested, not before.*

*The third paragraph's first sentence is in error. A dipole has extremes of reactance and a high radiation resistance and launches a wave from an element laying in one dimension. My square diagonal has unusually low and smooth reactance curves and a lower radiation resistance, and launches a wave from "an element" in two dimensions 90° apart, which gives a complex wave. The extra position on the switch would require a second set of tuners, as the electrical appearance to the feeders of a two-wire dipole is grossly different from a four-wire load. The pattern differences are now modest on the four wire systems, so that four patterns in place of two would be undesirable encumberance. I compared virtually every contact for a year between a 162-foot dipole and an equivalent 81-foot square square-diagonal antenna on 160 through 10 meters. One never performed "exactly as" the other.*

will perform exactly as a rotating dipole with the advantage of quick pattern change.

The "box diagonal" antenna is identical in pattern and polarization as above. It simply uses more hardware for the same result.

The "space dimension" antenna has the performance, pattern and polarization of a dipole that can be inclined at various angles to the horizontal and rotated in azimuth, by means of electrical switching. Similar comments apply to the "cube diagonal."

Since the direction of arrival and polarization of sky waves varies continuously, fast pattern and polarization changes can be a help for receiving. Such flexibility is not available in mechanically steered antennas. Since the arriving wave is generally linearily polarized it helps to have a receiving antenna that can be switched to circular polarization also. This can reduce fading markedly and does not require constant attention.

For transmitting purposes the problem is entirely different. Assuming the optimum sky path and expected polarization is not known in advance (which is generally the case), all that one can do is transmit under antenna conditions that yielded the strongest signal on reception. If polarization changes during the transmission there is no signal or feedback. Since polarization changes quite rapidly with skywaves there is little advantage for these antennas from a transmitting point of view.

For long distance work, on the average, the antenna that can put the most power into a given direction with the lowest possible angle will get through most often. There will always be "freak" propagation conditions where almost any kind of antenna can produce the strongest signal. Such things as "one directional" skip do certainly exist although they are extremely rare. Electro-magnetic path loss is usually identical at a given instant of time and polarization, in the two different directions.

**Peter Laakmann, WB6IOM**

*The box diagonal improved the "anti-fading" factor and the "DX in poor conditions" factor over the doublet, with similar feeders and tuners. These were delightful results of using more hardware.*

*If you are receiving my waves, which started on a triple polarization, instead of linear polarization, can you obtain as deep a fade on my signal? You may have thereby explained the reports I receive about less fading. Incidentally, what disturbs some engineers is that I observe reports of less dissipation per hop on the one hand, and on the other, feel that these systems have no great receiving advantages. I believe that the propagation damage happens to different styled waves differently in the ionosphere, (which itself is of peculiar polarizations), and after that fact, no receiving antenna can restore that extra dissipation. This is why I ask for transmitting, instead of receiving, tests through the ionosphere.*

*The antenna that yielded the strongest signal on reception (at my contact's location) was not the dipole—it was one of my multivector systems. Apparently I radiate a bushel-basket wave and overcome fading by being every which way at once. In optics a polarized filter can null light of singular polarization but it cannot null multi-polarized light.*

*Carl Mosley, an antenna manufacturer, in a technical talk at the Rochester Hamfest, said that quads (which have low-grade two dimensional vectorship) open the bands about a half hour earlier and close them about a half hour later than equal gain Yagis (which have high-grade one-dimensional vectorship). I think he was referring to poor propagation conditions, such as at band openings and closings. The low-grade two-dimensional waves of quads were many dB better than Yagis, but this difference does not remain when propagation is excellent. I observed the same confounding peculiarity on my continuous-spectrum, high-grade, two dimensional system, and then improved the effect by going to several three-dimensional systems. Comparative transmitting results are so unbelievably good, on poor skip, that after you have a three-dimensional system working, you will*

*be called a liar by those who never have tried one. When your DX contact is weak, all you have to ask is, "How does my signal strength compare with other American signals which you can hear now?" I suggest that you keep your tape recorder handy.*

**George A. H. Bonadio, W2WLR**

**Dear HR:**

Bill Orr's otherwise complete article on fm in the September issue glossed over one interesting point. This "footnote" is an effort to set the record straight.

Ham radio really missed the boat by not being the first service to pick up fm as an improvement over a-m. In the late 1930's, when fm was first suggested (magazines carried construction articles), putting an fm station on the air meant building a new kind of receiver (wide-band i-f, limiter and discriminator) and using it on either 5 or 2-1/2 meters. You can believe that no one was trampled in the rush, although a few brave souls did try it (and became believers).

In the next go-round, *narrow-band* fm was permitted on the more popular hf bands, but as Bill mentioned, it was used predominantly by hams trying to avoid bci (that's "broadcast-receiver interference" to the tv generation). Bill says that ". . . fm languished in the amateur bands . . . (it) was obtained by the flood of surplus a-m gear that invaded the market." That's just nice-guy Orr letting the hams off lightly. Look at the record. In the late 1940's every ham had an *a-m* receiver. To receive an fm signal he had to use slope detection (tune off to one side of the carrier frequency); you could copy an fm signal that way, but in doing so the fm signal took a 6-dB loss.

"So what?" you ask. Well, if you were using a 100-watt fm transmitter, you started out by taking an immediate 6-dB back seat to every 100-watt a-m rig on the band. Quite an incentive to change from a-m to fm!

Oh, but surely the hams realized that fm was better than a-m, for the reasons set forth in the technical articles of the time. Don't you believe it! Hams reasoned that their receivers told the true story, regardless of what the magazine articles said. They had observed a-m take out fm on their receivers; that was all the proof they needed.

Some manufacturers made fm detector/adapters that could be tied into an existing a-m receiver, to permit copy of an fm signal as it should be (and without the 6-dB disadvantage). But why invest in a gadget when your trusty station receiver had already proved that the *signal* (fm) was an inferior type? And thus, in 1939 and again in 1947, amateur radio booted a golden opportunity to grow up.

Then came ssb. On an a-m receiver it couldn't even be copied. But on a *cw* receiver (once you acquired the knack of tuning) sideband was loud and clear and had it all over a-m. And—here's the meat—every ham had an a-m/cw receiver. (How many—even now—have an a-m/fm receiver?) No ham had to buy anything to learn the superiority of sideband. Sure, he did have to learn how to use his receiver, but the little beauty was already paid for and all the owner needed was a few hits on the head and a little practice.

So we hams missed the boat by not picking up fm immediately, but we didn't boot sideband because our receivers could handle it. What do we do with the next great, or near-great, improvement that is offered to us?

Glad you asked. Here are a couple of suggestions:

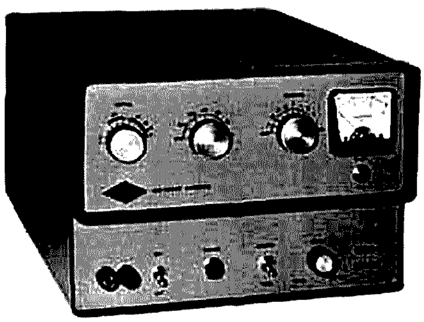
**1.** Understand the new thing. If you don't, or think you can't, find someone you trust and ask him to explain it to you. But until you *understand* the new thing, don't form an opinion! (That 6-dB disadvantage of fm disappeared with a suitable receiver, as did the "horrible, impossible, incompatible splatter of ssb.")

**2.** Don't brag about how quick amateur radio is to pick up new technical improvements. As Bill pointed out in his article, information on fm has been available since 1935 or so. It is a fact that hams are beginning to use fm at an increasing rate toward the end of the 1960's. This is alacrity?

**Byron Goodman, W1DX**
**East Hartford, Connecticut**

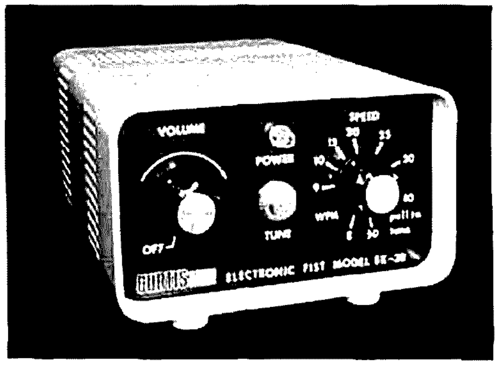
# new products

## new linear amplifier



Gonset has just announced a new linear amplifier for the amateur high-frequency bands—the GSB-201 Mark IV. This new grounded-grid amplifier is capable of the maximum legal input of 2000 watts PEP on 10, 15, 20, 40 and 80 meters and uses four husky carbon-anode 572B's along with a reliable solid-state full-wave bridge power supply. Both the high-voltage and bias supplies are built into the unit. Also included in the package is a universal antenna changeover relay that permits using the amplifier with either a transceiver, or with an independent transmitter and receiver. The cooling fan operates only while transmitting. $495 from your local dealer. Complete descriptive information is available from Gonset Division, Aerotron, Inc., P. O. Box 6527, Raleigh, North Carolina 27608.

## electronic keyer



Curtis Electro Devices has announced a new IC electronic keyer that features dot memory and instant character start. Speed range is 8 to 50 wpm. In addition to providing perfectly formed and spaced characters, the unit will provide semi-automatic operation with a straight key or "bug." Power supply, sidetone oscillator and speaker are built in. Sidetone pitch and volume controls are provided as well as momentary and locked "tune" switches.

A jack is provided for an external manual stand-by key. Output keying is switch selected for either solid state (—105V, 50 mA maximum) or read relay (optional). Cover is light blue wrinkle finish, with panels in dark green to match the popular Heath SB series. Other popular color combinations are available on special order for a slight additional cost. Complete, less paddle, $56.00; kit with assembled and tested card with connector, $25.00; relay option is $3.00 extra. Curtis Electro Devices, Box 4090, Mountain View, California 94040.

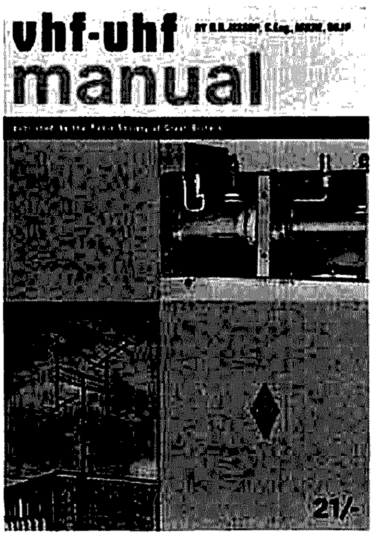
## worthington phasar-40 antenna

The new Worthington Phasar-40 is an electronically steerable directional vertical antenna array for 40 and 15 meters. The antenna system consists of two self-support-

ing quarter-wave radiators spaced either 34 or 68 feet. A remotely located control head allows selection of either a figure-eight or cardioid pattern. Rapid pattern switching allows the operator to null interfering amateur and foreign broadcast signals. The Phasar-40 provides a low angle of radiation for enhanced DX work. Properly installed, with an external ground system, the Phasar-40 will exhibit up to 4.5 dB over a single quarter-wave vertical, and up to 7 dB over a half-wave dipole.

Because of the relatively small diameter of the tube used for the vertical elements, the Phasar-40 is virtually invisible when viewed from any distance. Strength has not been compromised, however, and the antenna has been built to withstand 70 mph winds. $89.95 from Worthingham Electronics Company, P. O. Box 507, Warren, Michigan 48090.

## rsgb vhf manual




The Radio Society of Great Britain has published another excellent handbook for hams. It's the 245-page VHF-UHF Manual, written by G. R. Jessop, G6JP.

Perhaps the best way to point out its usefulness is to discuss the material it contains:

The first chapter is devoted to propagation, and should interest both the beginner and advanced vhf'er.

The chapter on tuned circuits and filters includes both design tables and formulas, and tested designs that are ready to copy. Of particular interest is the information on quarter-wave helical filters, which are little known, yet offer many advantages.

The chapter on receivers includes much general material, a complete hf/i-f receiver, an excellent discussion of noise, and many vacuum tube and semiconductor vhf/uhf converters and preamplifiers. One section of this chapter that will undoubtedly attract much attention is that devoted to tunnel diode amplifiers, including a practical 70-cm (432 MHz) preamp with a 3- to 4-dB noise figure. However, few would be likely to use one of these tricky beasts at 432 when easy-to-use fet circuits provide comparable performance.

Two 23-cm (1296 MHz) converters—one using transistors—and a unique crystal-controlled 13-cm (2304 MHz) converter are described. As is usually the case in RSGB publications, complete and detailed diagrams, layouts and tuning instructions make it easy to duplicate these converters.

The chapter on transmitters is as complete and interesting as that on receivers. Design information is provided on both tube and transistor equipment. Some descriptions that especially caught my eye were a true cavity 70-cm amplifier using a 4CX250, and simple 1/2-W, 2-meter and 70-cm solid-state transmitters that could be duplicated inexpensively in a few hours. Other chapters cover ssb, antennas and test equipment.

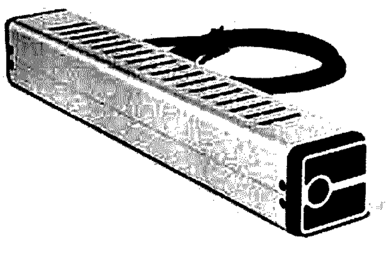
The fact that the book was written for English amateurs results in a few items of interest. The power limit is lower (150 W dc input), the popular band segments are slightly different (or in the case of 6 meters, quite different—they have 4 meters at 70 MHz), and some of the terms and components are unlike ours. However, these are minor and not liable to cause any problems. The RSGB *VHF-UHF Manual*, which is available for $3.75 postpaid from Communications Technology, Box 592, Amherst, New Hampshire 03031, belongs in the shack of every amateur interested in vhf and uhf.

# hot-carrier diodes

Motorola Semiconductor recently announced a hot-carrier diode designed primarily for uhf mixer and detector applications, but also suitable for fast switching circuits. The new hot-carrier diode, MBD101, is supplied in an inexpensive plastic package and sells for $.89 cents each in small quantities. The new MBD101 features low noise figure— 7 dB maximum at 1 GHz, as well as very low capacitance and high forward conductance. This new diode should be very interesting to amateurs who would like to try hot-carrier diodes but have been unable to obtain suitable devices.

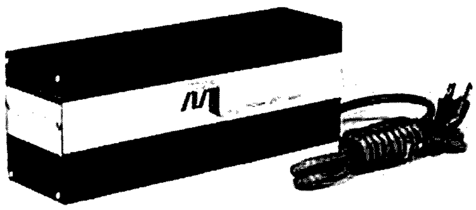
The Hewlett-Packard type 2800 hot-carrier diodes are now available for 90¢ each postpaid from H A L Devices, Box 365, Urbana, Illinois 61801. These are the diodes used in the three hot-carrier diode projects in the October 1969 issue of *ham radio.*

# low-cost lasers



If you're interested in trying your hand at laser communications, holography, interferometry or spectroscopy, suitable lasers are now available at very reasonable prices. The model 301 helium-neon gas laser from Quantum Physics, Inc., for example, is available in kit form or fully assembled. This laser produces 1 mW of uniphase radiation in the red-orange portion of the optical spectrum. The model 301 features a long-life cold-cathode plasma tube with an integral mirror mounter and integral power supply.

While the 1mW output power is considered well below the safe range, this laser is not a toy, but a sophisticated instrument that is suitable for many laser experiments. Because the laser chassis and power supply are prefabricated, the kit, which includes a prealigned laser tube and all mounting hardware, can easily be assembled in a few hours. Complete kit is $170; professionally assembled, $225. Larger 3 mW and 10 mW gas lasers available. Quantum Physics, Inc., 1295 Forgewood Avenue, Sunnyvale, California 94086.



Another entry into the low-cost laser market is the 0.5 mW Metrologic model 310, which was designed with the experimenter in mind. The model 310 operates in the orange-red portion of the visible spectrum, and emits about 0.5mW with a mean divergence of 0.5 milliradian. This instrument is ruggedly built and easy to take apart and reassemble. The price of the model 310 is $99.50. For more information, write to Metrologic Instruments, Inc., 140 Harding Avenue, Bellmawr, New Jersey 08030.

# low-resistance measurements

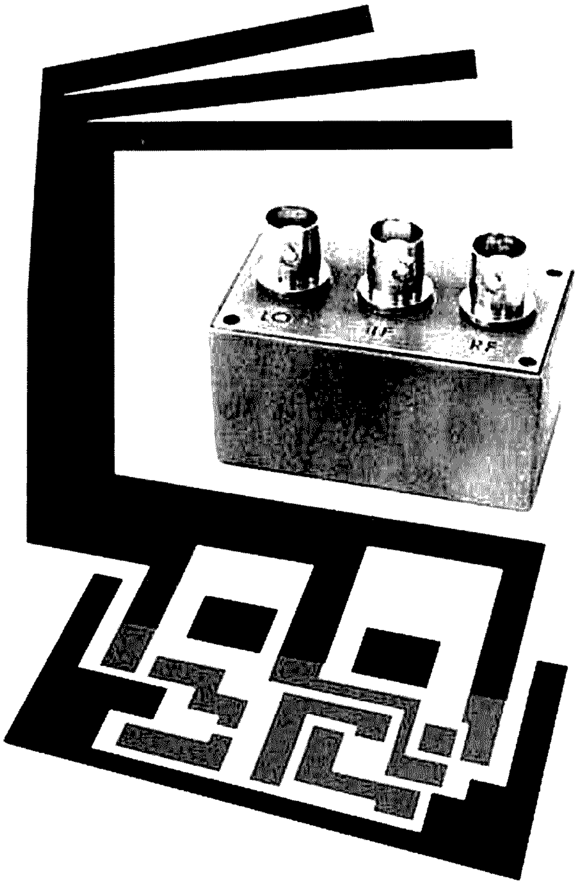
A new low-cost meter attachment capable of converting the Amphenol Millivolt Commander (and other 10-ohm centerscale meters) into a low-range ohmmeter has been announced. This new accessory is especially valuable for checking low value dropping resistors in solid-state supplies as well as for testing transformerless transistorized audio power amplifiers where the output resistance is typically in the 0.4- to 0.6-ohm range. The current is limited to 100 mA across 1 ohm, so the device under test is fully protected against accidental burnout. Price of the new Amphenol 870-3 milliohm meter attachment is $14.95. For more information, write to Amphenol Distributor Division, 2875 S. 25th Avenue, Broadview, Illinois 60153.

75 cents

focus on communications technology . . .

# ham radio magazine

MARCH 1970




# BROADBAND DOUBLE-BALANCED MODULATOR

## this month

march 1970
volume 3, number 3



**offices**

Greenville, New Hampshire 03048
Telephone: 603-878-1441

**ham radio** magazine is published monthly by Communications Technology, Inc. Greenville, New Hampshire 03048

Subscription rates, world wide: one year, $6.00, three years, $12.00. Second class postage paid at Greenville, N. H. 03048 and at additional mailing offices

Foreign subscription agents:
United Kingdom:
Radio Society of Great Britain,
35 Doughty Street, London WC1, England.

All European countries:
Eskil Persson, SM5CJD, Frotunagrand 1,
19400 Upplands Vasby, Sweden.



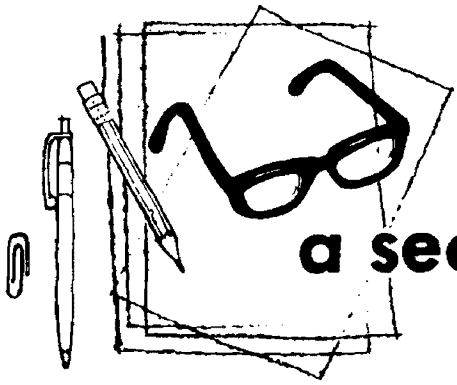
Printed by Wellesley Press, Inc. Wellesley, Massachusetts 02181, U.S.A.

**ham radio** is available to the blind and physically handicapped on magnetic tape from Science for the Blind, 221 Rock Hill Road, Bala Cynwyd, Pennsylvania 19440. Microfilm copies of current and back issues are available from University Microfilms, 313 N. First Street, Ann Arbor, Michigan 48103.

Postmaster: Please send form 3579 to **ham radio** magazine, Greenville, New Hampshire 03048.

**ham radio** magazine

# contents



# a second look

by jim fisk

**If your main interest** in ham radio is experimenting, you've probably tried all the published solid-state circuits plus some ideas of your own. Here are some suggestions for the experimenter who's run out of new fields to conquer in semiconductor applications.

Of the many devices on the market, few offer more interesting possibilities than the photodetectors. These emit a current in response to photon impingement, and have a spectral response determined by the energy band gap in their semiconductor material. Photodetectors with response in the infrared are especially interesting. In addition to their use as detectors in pulsed light-beam communications, they can be used in optical-electronic instruments for astronomy experiments.

Astronomers throughout the world are interested in obtaining data from amateurs as well as professional observers. Here's an opportunity to contribute to science and broaden your technical background as well.

Data is needed on the behavior of certain stars known to emit infrared radiation. Some of these stars are surrounded by gas and dust clouds. Much of their energy is absorbed by the dust and reradiated at longer wavelengths. Measurements at these wavelengths help clarify how the dust is distributed around the stars, thus providing clues as to their evolution.

A typical photometric system uses two in-line arrays of germanium bolometers with interference filters. To reduce noise equivalent power, detectors and filters are cryogenically cooled. Each detector looks at an optical telescope through a small-diameter focal-plane diaphragm. The filter for the first detector has a typical passband of 10-20 microns; that for the second detector has a passband of 25-50 microns. A dual-beam chopper, also cryogenically cooled, shifts the field of view of each detector through an arc equal to its angular diameter. A tracking system and finder telescope complete the assembly.

The system is aimed at an appropriate section of sky and locked onto a bright guide star offset from the infrared source, which may be invisible. Statistical fluctuations of the infrared source are measured, and the data is recorded on magnetic tape for later playback through a printer.

An elaborate setup such as this is beyond the means of most amateurs, but some interesting applications of semiconductors are implied. Some examples:

A photodetector array or photomultiplier could be placed at the focus of a telescope of moderate aperture, and detector output could be stored in a digital register driven by an a/d convertor. The output could be stored on tape or drive a counter directly. You could also probably improve on the clock drive used in most amateur telescopes with a system using IC op amps.

The rig would have to be used at the top of the highest mountain you could find because of atmospheric opacity at lower levels. Your first attempts might not yield much useful data, but don't be discouraged. Projects like this require much patience and equipment debugging. In any event, it's a pleasant way to spend a summer evening, especially if your best girl is along to help record data and maintain the log.

**Jim Fisk, W1DTY**
**editor**

## a word from the publisher

I am sure that by now you have noticed that *ham radio* has a new look this month. Like many other publications both large and small we have gone to what is known as "cold composition". By doing this we are able to take advantage of many new innovations in the graphics arts field.

As is usually the case with major changes, there are both advantages and shortcomings. One reader is sure to prefer our old type style, while another will think that our new one is an improvement. We will have more freedom to try new ideas, but last minute changes are now more difficult.

Of course, as this is written, I have not seen the very issue that you are now looking at. There may be parts which we will not like at first. You can be sure that we will be going over the results carefully to insure that we reach a new high in easy-reading and attractive appearance.

This issue also marks the first issue from our new printer: Wellesley Press. They have been chosen as experts in the periodical field. We felt they could add much to *ham radio* because of their great experience in the production of monthly magazines. Among the benefits should be more uniform quality and a great desire to give you, our readers, the magazine you want.

You will see a slight change in size. This has been done to permit us to take maximum advantage of the best high-speed printing equipment in use today. Our previous size was chosen because it was the same as *QST*. It turned out that there was a tremendous penalty involved and many excellent printers were virtually uninterested in working with it. This new size is a modern industry standard and gives us a marked advantage in holding down costs both to you and our advertisers.

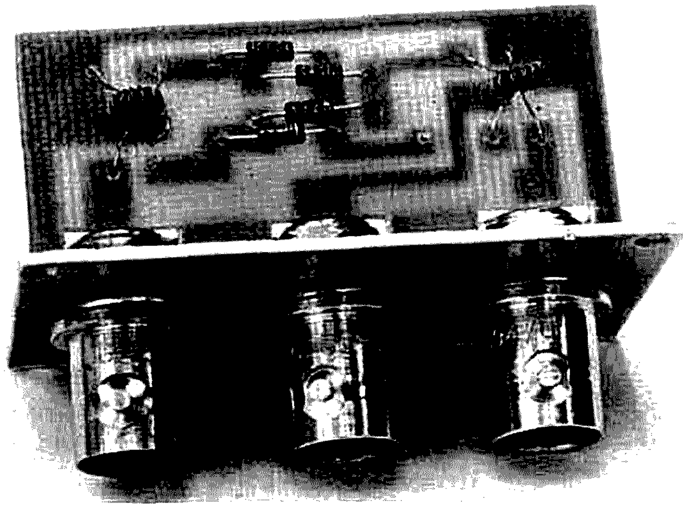
*Ham radio* is continuing to write one of the best success stories in the history of amateur radio publishing. 1969 saw our subscription list well more than double in size. This trend is continuing in 1970. We are still quite a way from being the biggest, but *ham radio* is a lot closer in size to some of its competitors than they would care to admit. Considering that we are now only two years old it certainly would seem that the ideas on which we based have been well proven.

## another topic

Our mailbox produced an interesting letter the other day which I thought a number of readers might like to share with us. It was from Mr. J. Cooper, G3DPS, General Secretary of the Royal Signals Amateur Radio Society in England.

This organization is now opening its membership to past and present members of the U.S. Army Signal Corps who have been attached to or worked in close liaison with Royal Signals. Many U.S. hams were sent to Britain during World War II and served in such a capacity. If you are eligible, this might well be an interesting chance to renew old friendships in other parts of the world. You can write to Mr. Cooper at the Royal Signals Amateur Radio Society, 15 Valley Road, Blandford Camp, Blanford Forum, Dorset, England.

**Skip Tenney, W1NLB**
**publisher**



# broadband double-balanced modulator

Practical construction details of a hot-carrier-diode double-balanced mixer that covers the range from 200 kHz to 250 MHz

William Ress, WA6NCT, Eimac Division of Varian,
301 Industrial Way, San Carlos, California 94070

Double-balanced ring modulators have been used since 1915, when they were developed by Bell Laboratories for carrier telephone systems. The earliest models were capable of good carrier suppression, but they suffered from high conversion losses because they used copper-oxide rectifiers. This, as well as excessive diode noise, limited their use to audio and low-frequency rf applications. With the improved semiconductor diodes that are available today, the double-balanced mixer circuit can be used in many communications applications that were formerly impossible; the home-built version presented here provides outstanding performance from 200 kHz to over 250 MHz.

Although double-balanced ring modulators require relatively high local-oscillator injection power, have some conversion loss, and must be followed by a low-noise amplifier, they have a number of operational advantages:

1. High port-to-port isolation (same as carrier suppression)
2. Wide dynamic range (large signal-handling capability)
3. Low intermodulation and cross modulation
4. Good noise figure
5. Reduction of spurious mixing products

It is also simple to build, easy to reproduce, fairly inexpensive and has a wide variety of applications in the radio communications field.

## diode mixers

The most simple diode mixer, of course, is the single diode type shown in **fig. 1.** This circuit is widely used in electronic equipment operating from audio through microwave – it's a good bet that you'll find at least one circuit like this in the amateur radio gear in your shack. However, this simple circuit has one very serious disadvantage: poor port-to-port isolation.

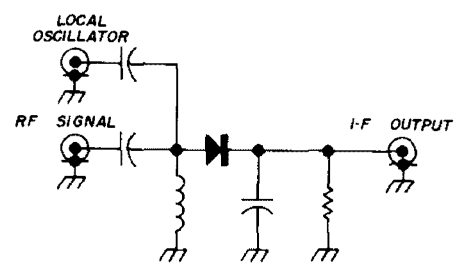



**fig. 1. Simple single-diode mixer lacks isolation between ports.**

The circuit of **fig. 2** was developed to improve isolation. Since transformer T1 is electrically balanced (center tapped), the local oscillator signal is split into two equal, but out-of-phase signals in the secondary, and cannot be induced into the primary. In the circuit of **fig. 3** the rf and local oscillator signals are isolated from the i-f output by virtue of the balanced secondary of transformer T2.

Further performance improvements can be obtained by going to the circuit of **fig. 4.** This four-diode double-balanced ring mixer allows energy to be exchanged on a full-wave cycle rather than half cycles as in the previous circuits and offers higher efficiency and lower conversion loss.

*Available from any Hewlett-Packard sales office; consult the Yellow Pages or write to Hewlett-Packard, 620 Page Mill Road, Palo Alto, California 94304. The matched quad, HPA 5082-2805 is $4.40. HPA 5082-2800 ($0.99 each) may be used but the matched quad provides better port-to-port isolation. These diodes are also available from HAL Devices, Box 365, Urbana, Illinois 61801.

## theory of operation

Consider the circuit of **fig. 4** with the rf input disconnected; with only local oscillator injection, there is no i-f output. When point A is negative, current flows through T1, diodes D2 and D3, and transformer T2, as shown by the arrows. Since the currents on each side of the center tap are 180° out of phase, they cancel, and there is no output. When point A is positive, current flows through T1, diodes D1 and D4, and transformer T2, again with no output.

If an rf signal is applied to T2, an output voltage appears across the i-f output terminals; the local-oscillator signal essentially switches the rf input voltage on and off. With high-conductance hot-carrier diodes, switching is nearly instantaneous and rectangular pulses controlled by rf signal amplitude appear across the i-f output port.

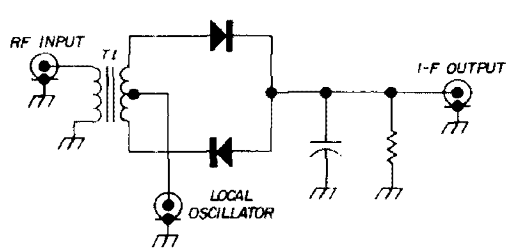



**fig. 2. The balanced transformer in this single-balanced mixer effectively isolates the rf port from the rest of the circuit.**

With a properly designed and constructed double-balanced mixer, carrier suppression on the order of 40 dB is not

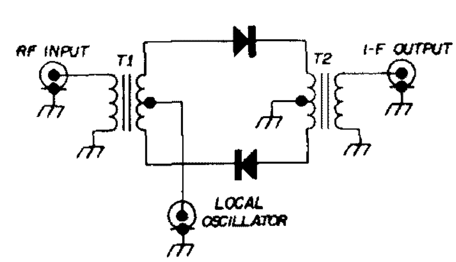



**fig. 3. Basic double-balanced mixer circuit.**

difficult to obtain. With this type of mixer, third-order distortion products are typically suppressed by 50 dB. Since the even harmonics are inherently suppressed by the double-balanced circuit, the only spurious signals that may give trouble are those created by odd-numbered harmonics.

## mixer components

The modern broadband balanced mixer is possible through the use of ferrite transformers and semiconductor diodes – diodes that exhibit high front-to-back ratios and ultra-fast switching times. Great strides in ferrite device technology has provided materials that operate efficiently from dc to microwave. With the proper ferrite and suitable windings, transformers can be made that will act as nearly purely resistive transformers over a wide range of frequencies.[1,2,3,4]

Although nearly any high-conductance diode will give adequate performance in a

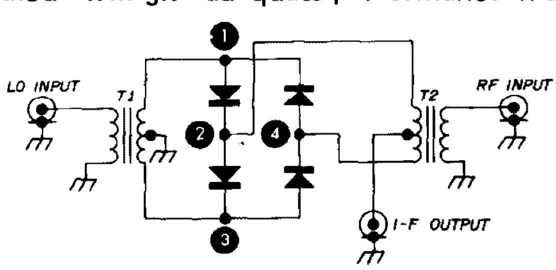



fig. 4. Double-balanced ring mixer circuit improves performance. Transformer construction is shown in fig. 6. Diodes are Hewlett-Packard 2800 series.

diode ring, many characteristics of the hot-carrier diode make it the ideal choice. To achieve electrical balance in the mixer for example, the diodes in the ring should have closely matched transfer characteristics – this is inherent in the fabrication of hot-carrier diodes.

For lowest mixer conversion loss, the ring diodes should have no forward resistance when conducting, and infinite resistance when turned off. The front-to-back ratios of several different diodes are listed in table 1. For efficient operation on the very-high and microwave frequencies, the mixer diodes should feature extremely fast switching speeds and contribute very little noise to the circuit. These requirements are best met by hot-carrier diodes.

The hot-carrier diodes currently on the market consist of a rectifying metal-to-semiconductor junction; n-type silicon in conjunction with evaporated gold, platinum, palladium or silver. In the hot-carrier diode, current conduction is based on majority carriers, and in normal operation the diode exhibits virtually no stored charge carriers. In practical terms

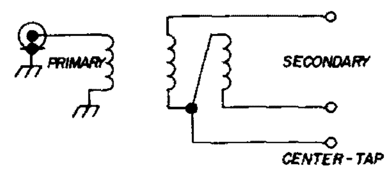



fig. 5. Schematic showing the connections of the trifilar windings.

this means that the reverse recovery time is very short: typically 100 picoseconds, more than four times faster than the fastest silicon junction diode. This results in more efficient signal rectification at vhf.

The construction of the hot-carrier diode results in uniform contact potential and uniform current distribution throughout the junction. In terms of operation, this means lower series resistance, lower contributed noise, higher power capability and greater resistance to transient pulse burnout.

## transformer construction

The transformers used in the practical double-balanced mixer shown in the photographs are wound on Indiana General Cf102-Q1 ferrite cores. These cores are available from Newark Electronics*. Each transformer consists of 12 trifilar turns of number 32 enameled wire; number 30 or 34 is also satisfactory. I experimented with a number of different cores and winding techniques to find a wideband design that could be

*Newark Electronics Corporation, 500 N. Pulaski Road, Chicago, Illinois 60624. Order catalogue number 59F1509, $1.20 each plus shipping. ($2.50 minimum mail order).

easily reproduced; the design described here performed the best.

To obtain the desired wideband performance, the coupling between windings must be as tight as possible. To obtain this, the three wires in each winding are twisted together: chuck two 2-foot lengths of number 32 wire into a hand drill (electrical drill if you're extremely careful); crank the drill until the wires have a reasonably tight twist. Then take this twisted pair and re-chuck it with the

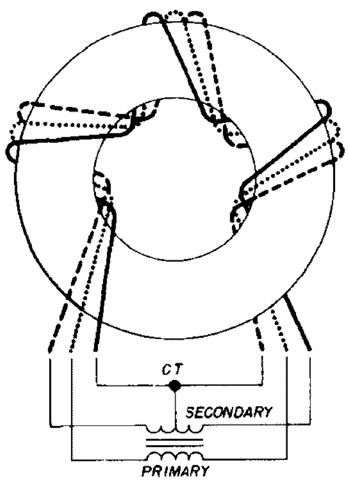



fig. 6. Toroid winding; wires are shown untwisted for clarity.

third wire and repeat the twisting process until you have a tight trifilar length of wire.

Each transformer consists of 12 turns of this trifilar wire on a CF102-Q1 ferrite core. A schematic of the complete transformer is shown in **fig. 5**; the pictorial diagram in **fig. 6** should explain the windings more fully. The windings must be connected properly, or the finished double-balanced mixer will not work. This can be simplified if you use different colored wire for each winding. I've been able to find the wire in two colors and even this is a big help.

Pick one set of wires for the primary. Wrap these two wires with a piece of tape to identify them and keep them out of your way. You should have four wires (two sets) left. Separate the two sets by checking for continuity with an ohmmeter. Take one wire from one set and twist it together with one wire from the remaining set; this is the secondary center tap. The two remaining wires will be the two outer ends of the secondary.

Now all the wire in the secondary will show continuity, and the two wires in the primary will be isolated from the secondary. The choice of wires for the primary and secondary is completely arbitrary – the only important thing to observe is the connection sequence.

To obtain the same wideband performance that I have achieved, the transformers must be duplicated. *If* you want to experiment, you might try some of the small cores from Indiana General in Q1, Q2 or Q3 material. Q3 material, for example, will improve the high-frequency performance at the expense of operation on the low-frequency end. Powdered-iron toroids should not be used because they will not operate over a very broadband frequency range.

## construction details

After the transformer windings have been selected you are all set to assemble the other parts of the cirucit. Here again I recommend following the layout I have developed; if at all possible, use the printed-circuit layout shown in **fig. 7.** If you use a different layout you won't duplicate my results. However, don't be afraid to try your own design – you may end up with better balance than I did. Just remember to use good vhf construction techniques: short leads and short ground returns.

The printed circuit is one area where

table 1. Front-to-back ratio of various diodes.

| diode type | forward resistance (ohms) | reverse resistance (ohms) | ratio |
|---|---|---|---|
| Copper oxide | 400 | 350k | 875 |
| Small-junction germanium (1N270) | 5 | 500k | 100k |
| Point-contact germanium (1N98) | 200 | 1M | 5k |
| Low-conductance silicon (1N457) | 50 | 2400M | 48M |
| High-conductance silicon (1N645) | 2.5 | 1200M | 480M |
| Hot-carrier (HPA 2800) | 1.5 | 3000M | 2000M |

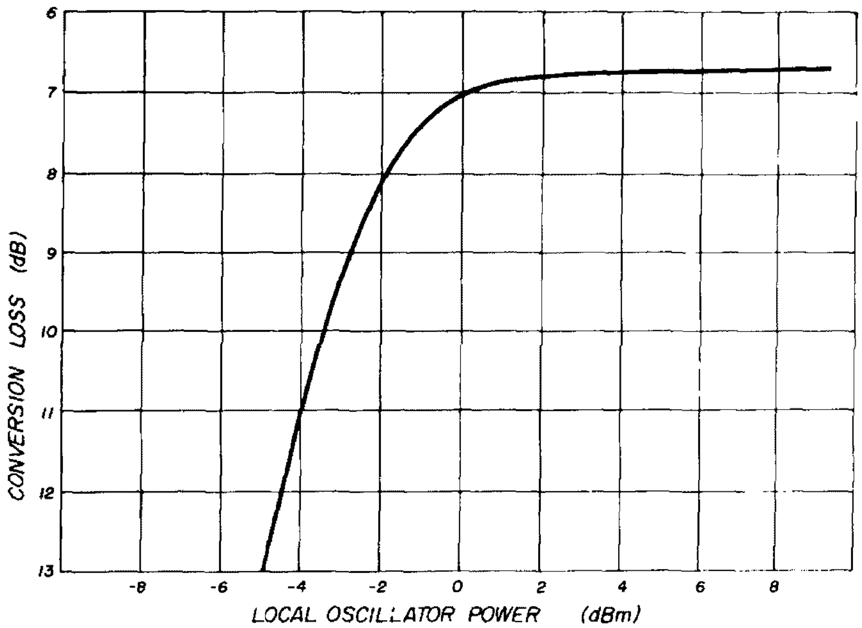



**Fig. 8. Conversion loss vs local oscillator power for this double-balanced mixer circuit.**

commercial manufacturers of these mixers use a touch of magic to obtain optimum balance. By using the stray capacitances associated with the circuit board and the components, it is possible to obtain nearly perfect electrical symmetry.

For proper operation, the completed mixer unit must be enclosed in a box that provides good rf shielding. In the unit shown in the photographs, I used a small cast-aluminum chassis manufactured by Pomona (model 2428). This enclosure sells for $1.50 at major electronics suppliers.

## applications

Probably the most important application of the double-balanced mixer in amateur equipment is as a frequency mixer. To obtain optimum performance as a mixer, three factors must be considered: local-oscillator power, conversion loss and the need for low-noise amplification of the i-f output.

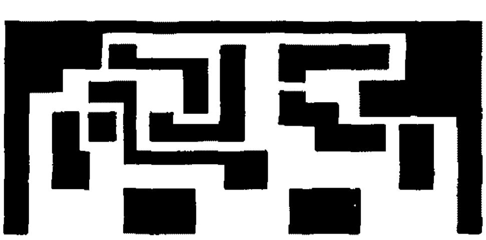


**Fig. 7. Printed-circuit board for the double-balanced mixer.**

The graph of fig. 8 shows conversion loss versus local oscillator power for this double-balanced mixer. This curve is typical of all passive mixers and shows that conversion loss decreases with increasing local-oscillator power up to approximately zero dBm (1 milliwatt or 0.22 volts across 50 ohms). Beyond zero dBm more local-oscillator power does little for conversion loss, but note how fast conversion loss rises as local-oscillator power drops below zero dBm.

Many active mixers work properly with as little as 0.1 mW (−10 dBm) of local oscillator injection, and are usually much more tolerant of variations in injection level. The relatively high local-oscillator power requirement of the double-balanced mixer is a disadvantage, particularly at uhf and microwave where it is harder to generate.

Conversion loss can be analyzed by putting an attenuator in front of the stage that follows the i-f output port. As an example, consider that you're using this mixer as a 144-MHz down-converter to 28 MHz, and you run the mixer's i-f output directly into the station receiver. Assume that the noise figure of the receiver is 10 dB at 28 MHz. Also assume that the mixer has a conversion loss of 7 dB.

The converter's noise figure is the

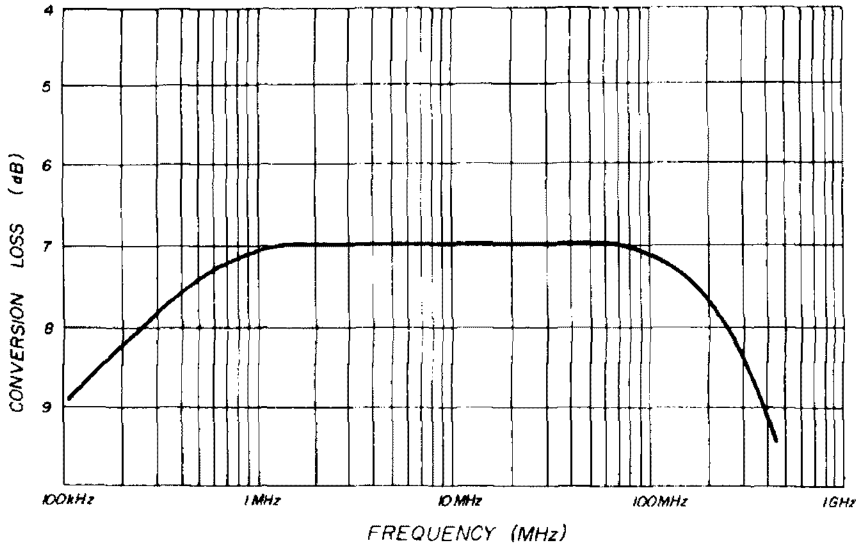



**fig. 9. Typical conversion loss for this double-balanced mixer over the range from 100 kHz to 400 MHz (local oscillator power +7 dBm, rf input - 5 dBm).**

receiver's noise figure *plus* the conversion loss, or 17 dB. This represents the noise contribution from both the mixer and receiver and assumes that the mixer is tuned to reject the image frequency. If the rf port is not tuned to reject the image the noise power in the image can add an additional 3 dB of noise; the converter would end up with an effective noise figure of 20 dB. This borders on the ridiculous for vhf converter applications but if we analyze the problem further we can find solutions that will change the mixer into a very useful vhf device.

To eliminate the images a filter is needed in the rf input. This can be a simple tuned circuit with a 50-ohm tap. The filter will eliminate noise contribution from the image but adds insertion loss. This must be added to the mixer's conversion loss. It's obvious that the filtering must have the lowest possible insertion loss. This can be accomplished with a wide bandwidth filter (same as low loaded Q). A good rule of thumb is to choose a filter with a bandwidth one-third the i-f output frequency.

We must also reduce the noise figure of the i-f. This is most easily done by adding low-noise amplification ahead of the receiver. A properly designed amplifier using transistors or fet's can yield noise figures as low as 1 dB at frequencies up to 60 MHz.

Let's take a look at an application using an rf input filter which has an

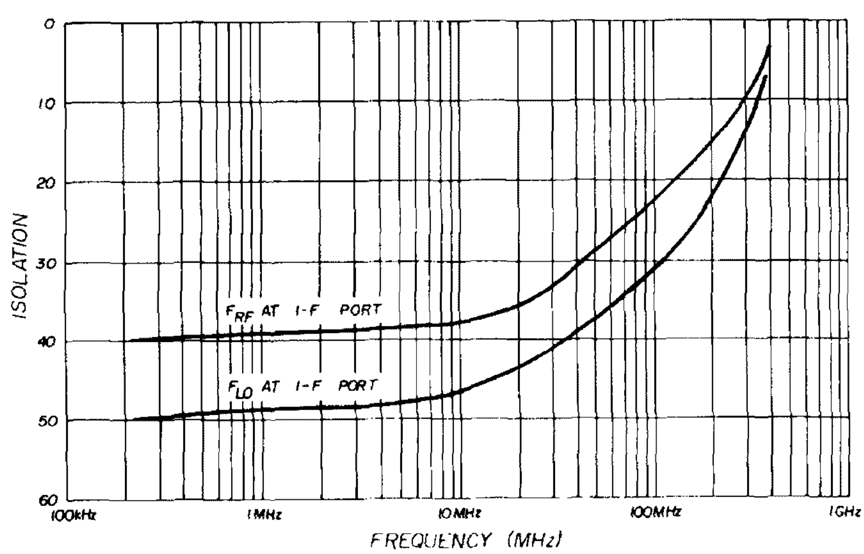



**fig. 10. Mixer balance in terms of isolation from i-f port (rf frequency at 0 dBm, local oscillator at +5 dBm).**

insertion loss of 0.2 dB and a low-noise amplifier ahead of the receiver which has a 2 dB noise figure. The mixer still has 7 dB conversion loss. We must add this to the insertion loss of the filter for a total of 7.2 dB; this 7.2 dB is added to the 2 dB noise figure of the i-f amplifier so the converter has an effective noise figure of 9.2 dB.

A front end with a 9.2 dB noise figure is useful for local ragchewing, fm repeater work and mobiling. For serious DX a low-noise preamplifier is required, but a 9.2 dB NF mixer can handle 1 milliwatt of signal before gain compression, cross modulation or intermodulation becomes a problem; only exotic active mixing schemes can accomplish this.

When you use a preamplifier ahead of the double-balanced mixer to reduce noise figure remember that the amplifier must have sufficient gain to overcome the mixer noise figure before you can realize the lower noise figure of the preamplifier. It's a good rule of thumb to design the preamplifier with at least 10 dB more gain than the noise figure of the following stage. In our example this would require

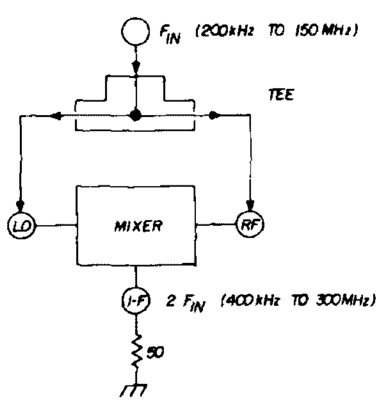



**fig. 11. The double-balanced mixer as a frequency doubler.**

19.2 dB preamplifier gain. Then the converter's noise figure would be set by the noise figure of the preamplifier.

I'd like to point out that the performance graphs for this mixer (**fig. 8, 9,** and **10**) compare closely with commercially available designs, although some mixers in the $100 class have improved port-to-port isolation and conversion losses as low as 6 dB.

I have discussed conversion loss of the double-balanced mixer, but have neglected noise figure. This is because the hot-carrier diodes contribute so little noise that it can't accurately be measured. Above about 1 GHz (1000 MHz) diode noise begins to become noticeable, and in the microwave region more exotic hot-carrier diodes are available that per-

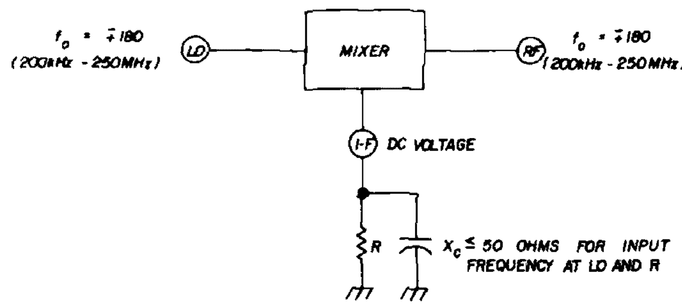



**fig. 12. Using the double-balanced mixer for a phase detector.**

form better than their more common silicon counterparts.

If you are using a diode mixer on 1296 MHz a properly selected hot-carrier device will offer a slight improvement in noise figure as compared to the old standby 1N21 series. Improved performance is even more noticeable on 2300 MHz and up, since the noise figure of hot-carrier diodes does not rise as fast with frequency as does the noise contribution of conventional point-contact and p-n junction devices.

In addition to its use as a simple frequency converter, the double balanced mixer is also useful for frequency doubling, phase detection, current-controlled attenuation, amplitude modulation, product detection and balanced modulation as shown in **figs. 11** through **17.**

## frequency doubler

The double-balanced ring modulator can be used as a broadband frequency doubler by simply applying the rf signal to both the local-oscillator and rf ports as shown in **fig. 11.** Since the sum and

difference frequencies will appear across the i-f port, the i-f output will be twice the rf input (since the difference frequency is zero).

## phase detector

When using the double-balanced modulator as a phase detector as shown in **fig. 12** one rf signal is applied to the local-oscillator terminals while the other rf signal is connected to the rf port. The dc signal available across the i-f port is zero when the two input signals are 90° out of phase; the dc voltage at the i-f port is maximum when the phase difference between the two signals is either zero or 180°.

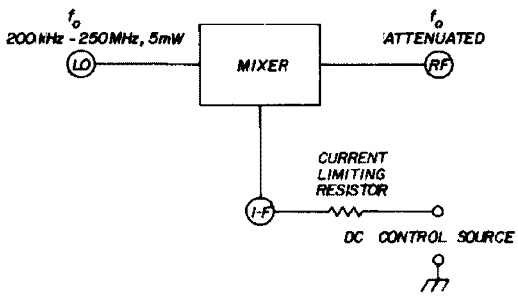



**fig. 13. Current-controlled attenuator; performance is plotted in fig. 14.**

## current-controlled attenuator

If you want to use the double-balanced mixer as a current-controlled attenuator, the rf input signal is connected to the local-oscillator port as shown in **fig. 13.** With no current input at the i-f port, the signal at the local oscillator port will appear greatly attenuated at the rf port. A curve of attenuation versus control current is shown in **fig. 14.** When using the mixer as a current-controlled attenuator, a current-limiting resistor should be connected in series with the i-f port to limit diode current to 40 mA. Appropriate values of resistance versus applied voltage are shown in **table 2.**

**table 2. Current-limiting resistance versus control voltage.**

| voltage | minimum resistance (ohms) |
|---|---|
| 1 | 27 |
| 5 | 150 |
| 10 | 270 |
| 50 | 1500 |
| 100 | 2700 |
| 500 | 15k |

If you refer to **fig. 4** you can see that a dc control voltage across the i-f terminals will cause two of the diodes in the ring to conduct. When sufficient dc current flows through the diodes they appear as very small resistors connecting the secondaries of T1 and T2 together, and any signal at the local-oscillator port will appear at the rf port with little attenuation. Varying the control current changes the resistance of the diodes, and hence, the magnitude of the output voltage.

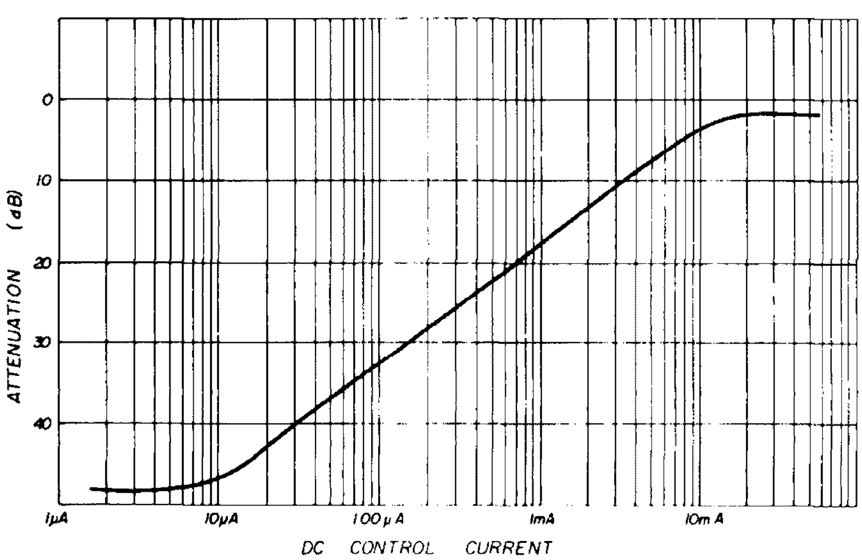



**fig. 14. Attenuation vs dc control current.**

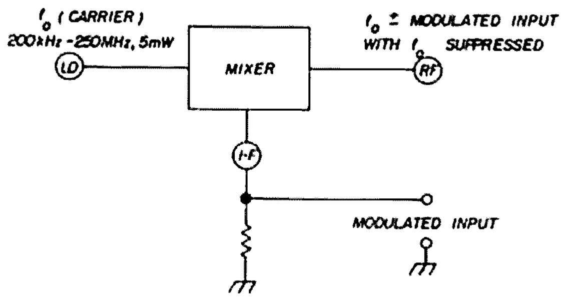



fig. 15. Balanced modulator.

## balanced modulator

To use this device as a balanced modulator, it is connected into the circuit as shown in fig. 15 with the rf signal (carrier) at the local-oscillator port, the modulating signal at the i-f port and the output signal across the rf port. The signal across the rf port consists of the local oscillator plus and minus the modulating signal with the local-oscillator (carrier) greatly attenuated.

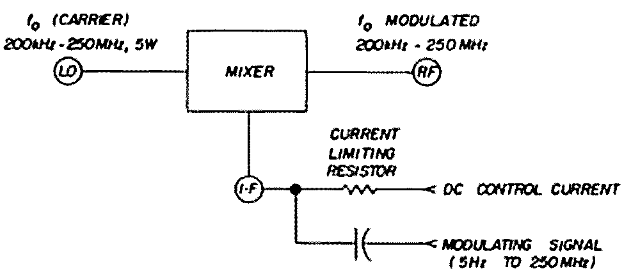



fig. 16. Amplitude modulator.

## amplitude modulator

To obtain amplitude modulation from the double-balanced mixer, the operations as a balanced modulator and current-controlled attenuator are combined as shown in fig. 16. A modulating signal containing both ac and dc components is applied to the i-f port. The ac components will produce sidebands and the dc component will vary the amplitude of the carrier appearing at the rf port. For 100 percent modulation the modulating signal should be about 200 mV rms and the dc control current should be approximately 4 mA.

## product detector

This is simply a mixer that has its i-f output in the audio range. A suitable circuit is shown in fig. 17. The double-balanced ring mixer is particularly useful in this application because of its very low intermodulation performance and large dynamic range.

### two-meter converter

The two-meter converter shown in fig. 18 is based on the hot-carrier-diode double – balanced mixer shown earlier. This converter has all the design features that should be considered when using an hcd mixer in the converter, including an input filter, low-noise i-f amplifier and a spectrally clean local oscillator.

The two-meter converter shown in the photo has a noise figure of 9 dB; and gain can be adjusted from zero to 20 dB. Main image rejection (84 to 88 MHz in this case) is 30 dB; all other images are down at least 60 dB. Local oscillator leakage at the input and output ports is 500 microvolts. The gain compression

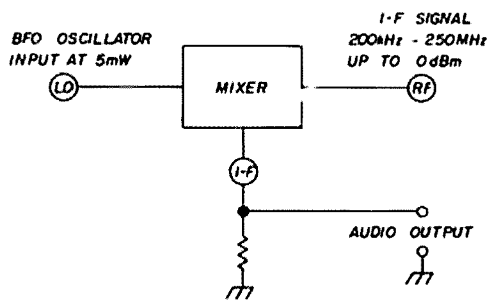
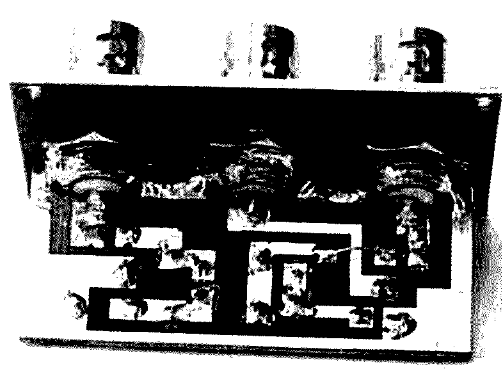
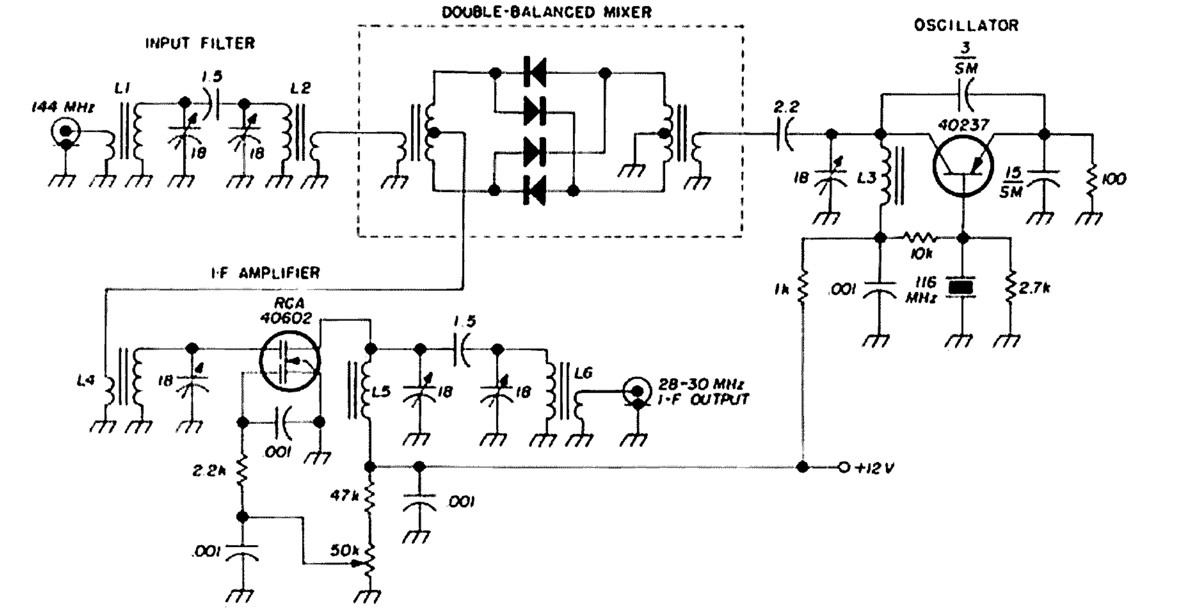



fig. 17. Using the double-balanced mixer as a product detector.

Construction of the hot-carrier-diode double-balanced mixer showing the circuit side of the printed-circuit board.






L1, L2 Primary is 10 turns no. 24 on Micrometals* T30-10 toroidal core; secondary is 4 turns no. 24 on cold end of primary

L3 7 turns no. 26 on Micrometals T30-22 core
L4 24 turns no. 28 on Micrometals T30-6 core
L5, L6 24 turns no. 28 on Micrometals T30-6 core
Secondary of L6 consists of 3 turns no. 28

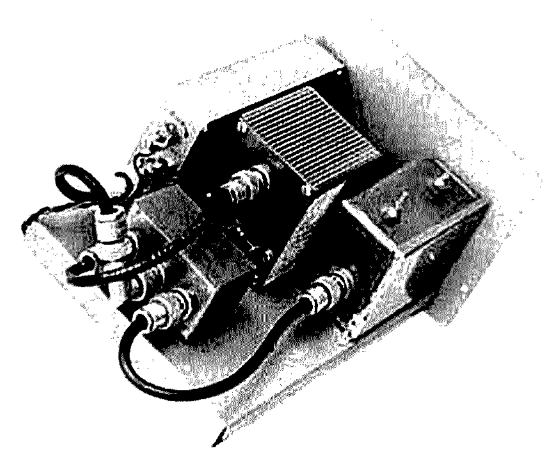
**fig. 18. High-performance two-meter converter is based on the double-balanced mixer package.**

point the point where the output departs from linear change relative to the input change is 1 volt rms.

In the converter shown in the photo, each of the main components was built into a separate chassis. This improves shielding between stages and facilitates experimentation with different converter configurations.

The hot-carrier-diode double-balanced mixer used in this converter has dramatically demonstrated to me the ability of a passive mixer to offer high dynamic range and resistance to overload, desensitization, cross modulation and intermodulation, while providing a respectable and useable noise figure.

**Two-meter converter using the double-balanced mixer. The mixer is on the rear of the chassis; in front of it, from left to right, are the low-noise 30-MHz i-f amplifier, 116-MHz local oscillator and 144-MHz input filter.**



## references

1. R. Turrin, W2IMU, "Broadband Balun Transformers," *QST,* August, 1964, p. 33.
2. C. Ruthroff, "Some Broadband Transformers," *Proceedings of the IRE*, August, 1959, p. 1337.
3. W. Orr, W6SAI, "Broadband Antenna Baluns," *ham radio*, June, 1968, p. 6.
4. D. Thorpe, "The Magic Tee," *73*, September, 1969, p. 14.
5. E. Pappenfus, et al, "Single Sideband Principles and Circuits," Chapter 5, McGraw-Hill, New York, 1964.
6. J. Fisk, W1DTY, "Double-Balanced Mixers," *ham radio*, March, 1968, p. 48.
7. "The Hot-Carrier Diodes: Theory, Design and Application," Hewlett-Packard Application Note 907, 15 May, 1967.
8. H. Sorensen, "Using the Hot-Carrier Diode," *Hewlett-Packard Journal*, December, 1965.

* Micrometal toroidal cores are available from Amidon Associates, 12033 Otsego Street, North Hollywood, California 91607. Package of six cores for the two-meter converter is $2.75 postpaid.

ham radio

# compact dual-band antennas

**Simple but effective antenna systems for city-lot dimensions or portable use**

William I. Orr, W6SAI, Amateur Service Department, Eimac Division of Varian, San Carlos, California 94070

It's hard to "be loud" when your antenna is on a city lot. Power lines and apartment buildings make it almost impossible to put up a full-sized antenna. It is possible, however, to reduce the length of a dipole by one-half using the folding technique. A shorter dipole combined with a folded dipole is an efficient two-band antenna system that can be erected in a restricted space. Such an arrangement is also useful for portable or Field Day work. Versions of this system are discussed in this article.

## basic concepts

Two dipoles may be connected in parallel at a current loop as shown in fig. 1, provided an even-harmonic relationship exists between them. End separation of both antennas need be only a few inches. Although a small amount of detuning will exist, the usual formulas for dipole length apply.

Various combinations are practical for two-band systems: 160/80, 80/40, 40/20, 20/10, and so on. The 40/20-meter combination will work on 15 meters if the 40-meter antenna is operated on its third harmonic. An 80/20 or 80/10-meter system can be built (using the even-harmonic rule) with a reasonably low standing wave ratio on the transmission line. As with any system using an unbalanced transmission line feeding a balanced antenna, a balun should be used to preserve antenna pattern and to avoid feed problems.[1]

## folded half-wave radiator

The length of this simple dual-band system can be reduced by folding the lower-frequency antenna back on itself. A three-wire antenna is shown in fig. 2. The feed point is connected to one of the outside pairs of wires and also to the inner pair. The two outside wires are jumpered at their far ends; they are the elements of the low-frequency dipole.

Folding the antenna has a minimum effect on its resonant frequency. If you'd like to refine the resonant frequency adjustment, compensation may be made

in the manner shown in fig. **3.** The low-frequency dipole should be trimmed for the low-frequency end of the band. The resonant frequency can then be raised by moving an adjustable jumper across the wires at the end. If the jumpers are adjusted in unison, resonant frequency may be varied over several-hundred kHz.

The higher-frequency dipole (center wire) will be unaffected by this adjustment. It may be adjusted by changing its length until resonance is achieved.

## system bandwidth

The bandwidth of any antenna may be defined in terms of the allowable standing wave ratio on its transmission line. Beam antennas with close-spaced parasitic elements have very low radiation resistance and limited bandwidth. If they are operated at an swr much higher than 2:1, forward field and front-to-back ratio will deteriorate rapidly at frequencies only a few percent from resonance. Simple dipoles, on the other hand, can operate over a much wider frequency range. This is because there's no problem involving phase and reactance relationships between elements as with parasitic beams.

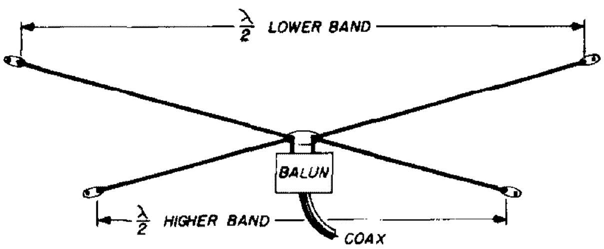



fig. 1. Basic dual-band antenna. Dipoles are parallel-connected; an even-harmonic relationship must exist between them.

Equipment limitation is probably the most important factor that affects allowable transmission-line swr. Many commercial amateur transmitters and transceivers have limitations of 2:1 for standing wave ratio to avoid ruining output-circuit components (usually a pi network). It is therefore prudent to operate such equipment into antenna transmission lines with an swr of 2:1 or less. Let's see how the dual-band radiator measures up to this criterion.

## height above ground

Taking an swr of 2:1 as par, the plot of fig. 4 shows the swr of the dual-band dipole when operated at optimum height above ground. The data for these curves was taken at the end of a 100-foot length of 50-ohm coaxial cable.

The 40-meter dipole has greater bandwidth: more than 400 kHz, with an swr of 2:1 or less. The 80-meter folded dipole's bandwidth is about 75 kHz over the same swr range. This indicates that the 80-meter antenna height is more important, in terms of swr, than that of the 40-meter antenna. I raised and

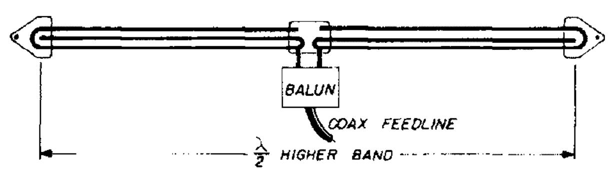



fig. 2. Dual-band system using the folding technique. Center wire is the higher-frequency dipole; the two outer wires comprise the lower-frequency dipole. Interaction is negligible.

lowered my 40-meter antenna and observed its swr across the band. It performed satisfactorily at heights above 20 feet or so, with a rather small change in swr. The swr reached a broad optimum at a height of 30 feet and again at 60 feet. Highest swr was at about 45 feet.

The 80-meter antenna is a different breed of cat. Because of folding, the 80-meter dipole's radiation resistance is lower. It's about 60 percent of the usual measured value at all heights above ground. The 80-meter antenna plot of fig. 4 occurred at about 50 feet and remained reasonably constant down to 40 feet. Below this height, minimum swr increased rapidly, tending to decrease the over-all bandwidth.

Accepting these facts of life, I finally mounted the antenna so that the flat top was about 45 feet high. The 40-meter-band swr wasn't as good as shown in fig. 4; however, it remained below 2:1 from

7.1 to 7.3 MHz It was still acceptable at 7.0 MHz, as no equipment-loading problems were encountered at this frequency. The 80-meter swr values were as shown in fig. 4.

## feed system

The transmission line must be decoupled from the antenna to obtain lowest swr. Decoupling will keep the line from radiating. A sure way to create transmission-line problems is to attach an unbalanced coaxial line to a balanced antenna. The reason is that current flows down the outside of the line instead of being confined to the inner conductor. A balun placed at the dipole feed point will decouple the line's outer shield from antenna currents. The balun is shown in fig. 5. It's mounted directly at the center antenna insulator.

## balun construction

The balun consists of three trifilar windings of no. 14 enamelled copper

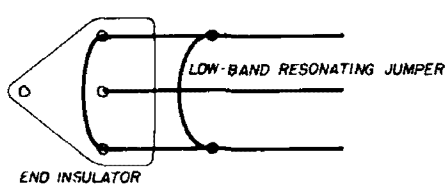



**fig. 3. End insulator and resonating jumper. Each side of the assembly should be of equal length; jumpers should be moved in unison.**

wire. Each winding consists of 8 turns. The coils are wound over a length of high-Q ferrite rod, ½-inch in diameter.* Nick the rod with a file around its circumference at the desired length, then break it with a sharp blow. Connect the ends of the windings as shown in fig. 5.

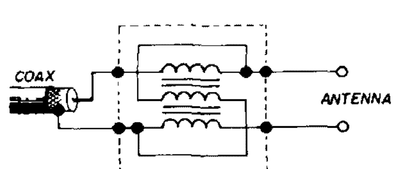



**fig. 5. Balun for optimum performance. Inductances are wound over ferrite cores; see text.**

Note that the coax shield is connected to the *outer* winding and also to the opposite end of the center balanced winding.

## transmitter loading

The RG-8/U feed line (or RG-58/U if power is below 250 watts PEP) should drop vertically as far as possible. The line then may be run horizontally when near ground level. Since a relatively high swr exists, it may be necessary to vary line

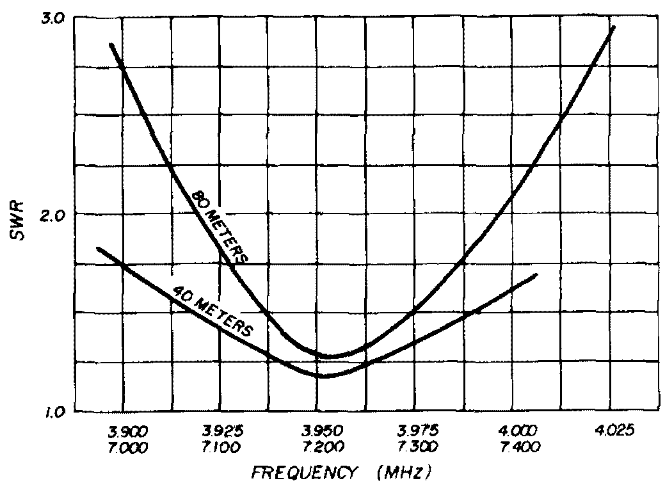



**fig. 4. Standing wave ratio versus frequency for the compact dual-band dipole. Curves were optimized for antenna height.**

length to obtain optimum transmitter loading. This doesn't change the swr; it merely provides conditions for a better impedance match at the transmitter end of the line. Make a couple of ten-foot lengths of line with appropriate connectors. Try inserting one or both sections into the main transmission line until the transmitter loads properly. If loading difficulty still persists, a longer line section may be necessary to obtain proper loading on both bands. Line length isn't nearly as critical as it may seem — I mention it only because loading difficulties might develop.

## dual-dipole construction

Construction is simple. The end and

*Indiana General CF-503 rod, available from Newark Electronics. Catalog part no. 59F-1521.

center insulators may be made of 3/8-inch plywood squares about six inches long. For outdoor use the insulators should be treated with spar varnish to make them waterproof.* Holes for the antenna wires should be about two inches apart. String the wires, then stretch the system between two supports about waist high. You'll notice that separators will be needed between the wires, spaced at about 5-foot intervals. You can make these from short pieces of plastic rod.

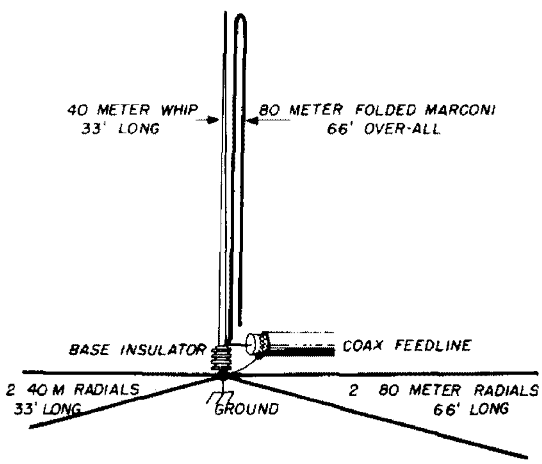



**fig. 6. Compact dual-band Marconi antenna. Ground system is important; a ground rod was used in addition to the radials, which can be oriented at random.**

Drill the separators to accept the wire, then thread the separators onto the wires. Secure the separators with small-diameter wire ties.

The first antenna of this type that I built had untreated insulators and separators. They lasted about a year, then succumbed to a combination of weather and birds. This construction isn't recommended except for temporary installations.

Commercial versions of this compact antenna system are available. When used with a balun, performance will be as described here. Again, I'd like to emphasize that this antenna, when operated without a balun, will lead to unusual or puzzling operating conditions. Play it safe and do the job right.

Typical antenna dimensions are given in table 1.

*Another time-honored method of waterproofing wooden insulators is to boil them in paraffin. A one-pound block of paraffin is less expensive than a quarter-pint of spar varnish. *Editor.*

## dual-bank Marconi antenna

The parallel-feed system may be adapted to Marconi antennas as well as to dipoles. With a Marconi, a ground system is required for proper operation (**fig. 6**). The random "water-pipe" ground is not recommended. Two or three radials should be used at the ground connection in a Marconi antenna installation. Two ¼-wavelength radials for each band will be adequate. The radials may be of insulated wire; they don't necessarily have to form the spokes of a wheel from the ground connection. They can be fastened to fences or any handy anchoring device.

The Marconi is usually in the form of a base-supported whip. The easiest way to erect a Marconi antenna for two-band operation is to use a ¼-wave whip made of tubular material for the higher-frequency band, which acts as a support for a folded-wire section cut for the lower-frequency band. Insulators such as those used for tv lead-in may be installed on the whip to support the folded-wire antenna The folded-wire antenna will require insulated spreaders, as discussed previously.

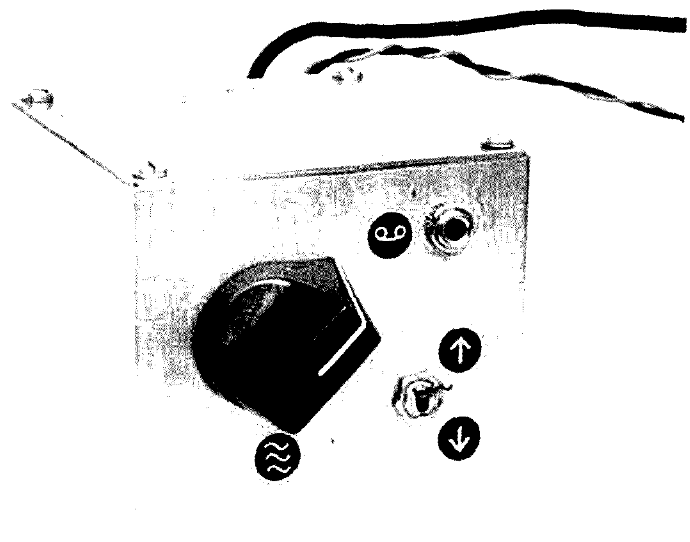
Radiation resistance of the whip antenna is lower than that of the dipole system. However, height above ground has less effect on the whip's bandwidth. A typical 80/40-meter Marconi, for example, has an swr of 2:1 or less across the 40-meter band. An 80-meter system has a bandwidth of about 80kHz.

An added benefit of the vertical Marconi is a low radiation angle, which is good for DX work. Best results will be obtained only if a good radial system is used.

### references

1. William I. Orr, W6SAI, "Broadband Antenna Baluns", *ham radio*, June, 1968, p. 6.

**ham radio**



# tunable peak-notch audio filter

**Solid-state circuits featuring the twin-T network - useful in test equipment or for improving receiver selectivity**

John H. Schultz, 40 Rossie Street, Mystic, Connecticut

Numerous filters have been described that provide selectivity at audio frequencies. The usefulness of many such filters is somewhat limited for several reasons. Some provide only one function: *either* peaking or notching. Others, such as the Selectoject[1], provide both functions over a tunable audio-frequency range. However, these haven't been adapted to solid-state circuits, nor have their RC networks been optimized for frequency selective peaking or notching. Still other designs have appeared from time-to-time, but are rather complex and expensive because of the several cascaded tuned circuits required to obtain good selectivity characteristics.[2,3,4]

A tunable audio frequency peaking and notching (rejection) filter is one of the most useful accessories you can have. Besides providing effective audio selectivity for ssb or cw reception, the audio filter is a useful adjunct in conducting

*harmonic- and intermodulation-distortion* measurements and in measuring frequency in the af range.[5]

The audio-frequency-selective network described in this article is not new, although it has been overlooked in recent years. What is new, however, is the manner in which the network is employed to provide a selectable frequency-peaking or frequency-notching function, tunable over a wide portion of the audio-frequency spectrum, using only resistors and capacitors. The basic network is discussed as well as some practical circuits in which it can be used. Many other applications for the network will probably suggest themselves.

## basic network

The basic frequency-selective network is shown in **fig. 1A**. Called a twin-T network, its theory is discussed in electronic texts, so I won't repeat it here. The network is equivalent to the Wien bridge. It passes all frequencies except one, where the relationship

$$\text{rejection frequency} \cong \frac{1}{2RC}$$

is met. If the capacitor values are fixed and the resistors are variable, a tunable notch, or rejection, circuit results. By rearranging the network, a single-frequency peaking circuit results **(fig. 1B)**. Thus, with appropriate switching, a selectable notching or peaking filter can be designed.

As mentioned before, the basic network has been around for some time and has been used by amateurs with different degrees of success. Its unsuccessful use can be traced to several factors. First, the network can't be loaded on either its input or output; it must be used in high-impedance circuits. Secondly, the components (mainly the capacitors) must have low internal impedance and be closely matched. The usual assortment of capacitors in the junkbox simply will not work. Finally, to be truly tunable, all three resistor legs must be variable, not just two.

## circuit applications

Fig. 2 shows the basic network, switchable for either peaking or notching, in a low-level transistor amplifier circuit. The amplifier can be used between the stages of another amplifier where audio selectivity is desired, or it can be used to drive a pair of high-impedance headphones directly from a receiver or test instrument.

Network component values were chosen on the basis of being readily available in the correct ratios and providing a reasonably wide tuning range. Coverage is from about 300 to 10,000

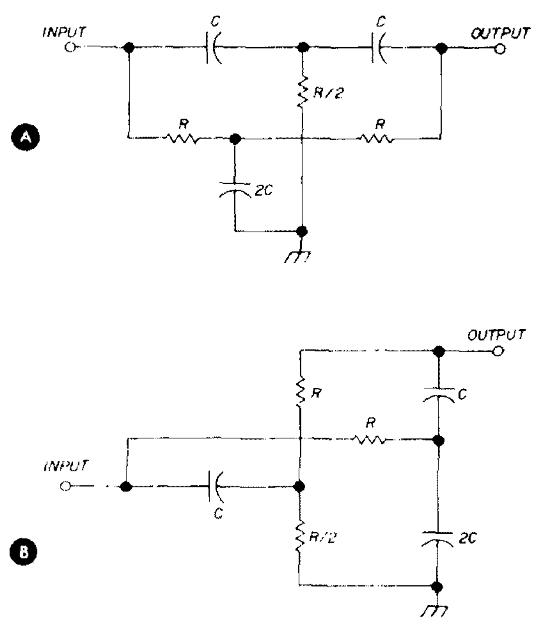



**fig. 1.** Basic twin-T notch circuit, A. An alternate arrangement for peaking is shown in B.

Hz, which should satisfy most requirements. The frequency can be made lower by increasing both resistor and capacitor values.

The impedance across the input transistor base-emitter junction isolates the network from loading effects. A feedback path is provided from the output stages to one leg of the filter. This raises the terminating resistance presented to the filter, which improves selectivity characteristics.

The selective network in a complete, self-contained audio unit using an IC is

shown in fig. 3. It can be plugged into the headphone jack on a piece of equipment. The input transformer's secondary-winding impedance prevents network

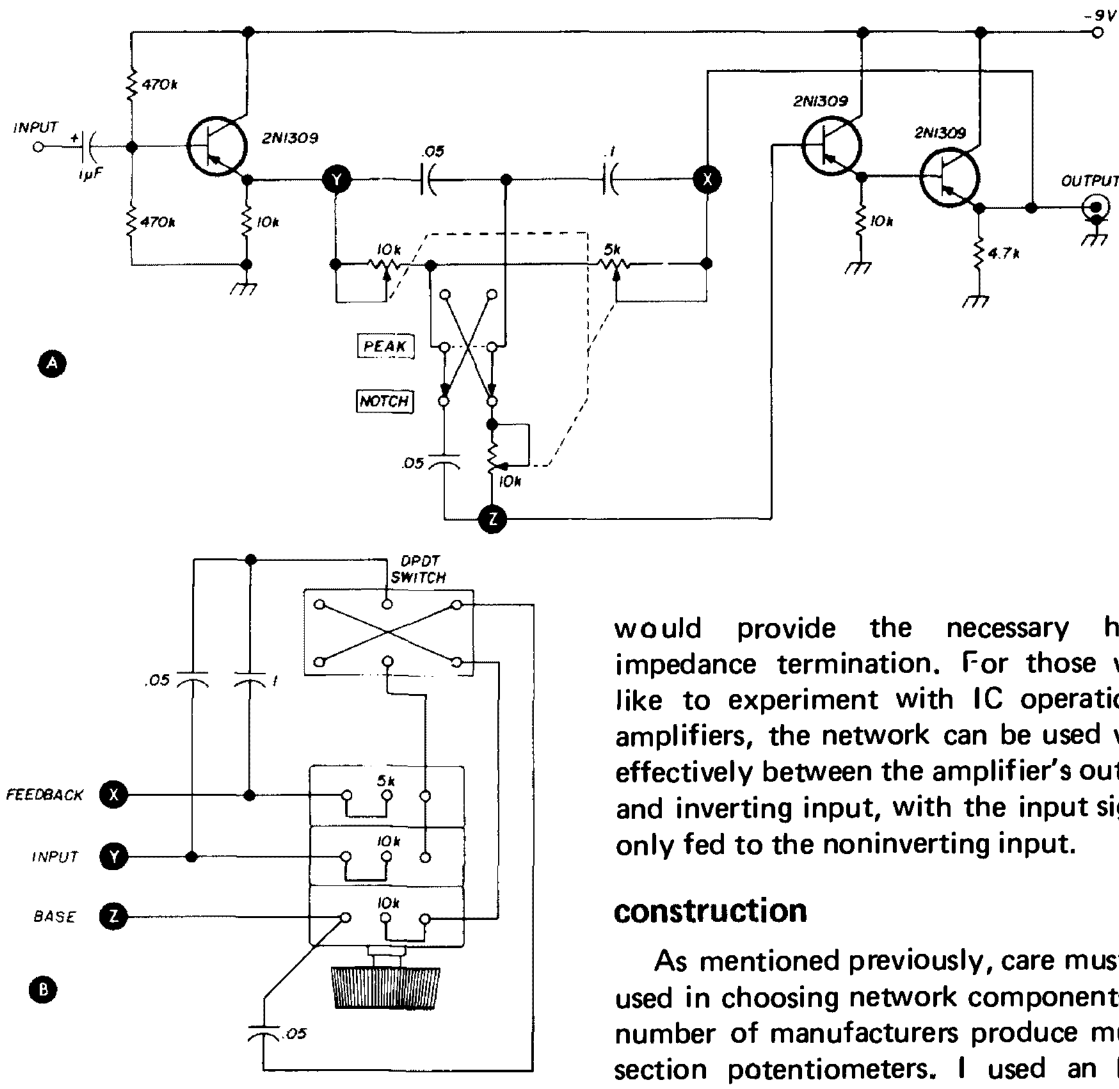



fig. 2. Variable-frequency peak-notch filter with feedback, A, that can drive high-impedance headphones or an audio amplifier. B shows potentiometer and switch wiring.

loading. Transformers with much low impedances should not be used. The audio amplifier was designed originally for use with a crystal phone pickup; thus it provides a high terminating impedance for the network. A 9-volt transistor battery can be used for power; however, if the unit is used at its maximum ½-watt power output for an extended period, a heavier-duty power supply would be required.

The basic network can be used in other circuits as well. For example, in circuits employing fet's, these devices would provide the necessary high-impedance termination. For those who like to experiment with IC operational amplifiers, the network can be used very effectively between the amplifier's output and inverting input, with the input signal only fed to the noninverting input.

## construction

As mentioned previously, care must be used in choosing network components. A number of manufacturers produce multi-section potentiometers. I used an IRC three-section pot with linear tapers. Capacitors were Sprague 10-percent tolerance tantalum.* Similar low-loss components may be used, of course.

The photo shows how the circuit of fig. 2 is assembled in a small 2-inch-cube Sta-loc enclosure. This enclosure allows all six sides to be separated and is convenient for compact construction. A small minibox can also be used. The transistor circuit is assembled on the

*Potentiometer is Allied stock no. 46F1892C. Specify 45D103, Md103, Md50218 for each section. Capacitors are Allied stock no. 43F4923 (0.47 μF) and 43F4926 (0.1 μF). Allied Radio, 100 N. Western Avenue, Chicago, Illinois 60680.

vector board mounted next to the potentiometer.

Power was taken from the unit with which the filter was used, but a slightly larger enclosure would accommodate a 9-volt transistor battery.

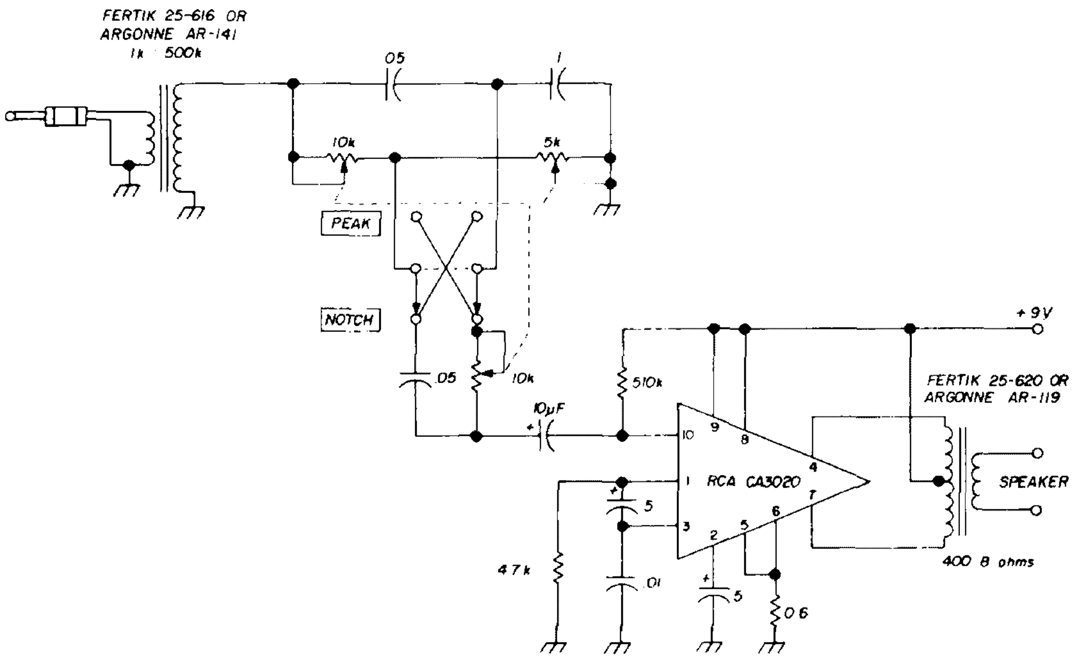



fig. 3. Twin-T filter combined with an IC. Unit provides about ½ W output and can be plugged into the headphone jack on your receiver.

## summary

This simple network can provide a remarkable degree of selectivity in circuits where a gradual frequency rolloff is acceptable. It doesn't require complex arrangements using toroids, multiple tuned circuits, etc. Peaking and notching response of the circuit of **fig. 2** is depicted in **fig. 4.** Filter response is narrower as frequency setting increases.

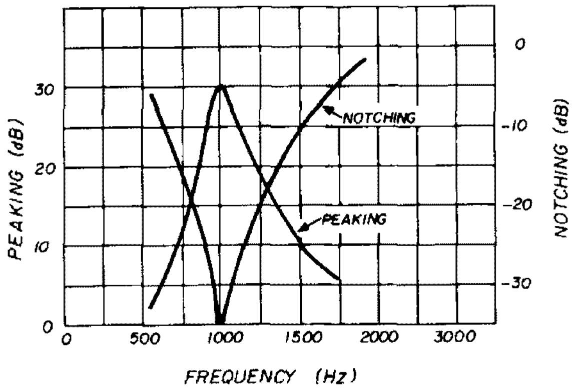



fig. 4. Typical response of the circuit in fig. 2. Response broadens somewhat at lower frequencies and is narrower at higher-frequency settings.

Despite the comments made earlier about not loading the network, you might want to try the basic network in a headphone circuit. I couldn't resist trying it with 400-ohm phones. Results were moderately successful but hardly equal to those using the transistor circuit. Although using the network alone can't really be recommended, it does work to some degree and provides some selectivity as a passive circuit.

## references

1. ARRL staff, "Radio Amateurs Handbook", 1957 edition, p. 132.
2. Louis N. Anciaux, WB6NMK, "A Solid-State Audio Filter", *QST*, December, 1968, p. 35.
3. Donald J. Sommer, WA7FBO, "Selectable Bandwidth Filter", *QST*, August, 1968, p. 47.
4. Harry J. Gensler, Jr., K8OCO, "The OCO Audio Filter", *QST*, January, 1962, p. 16.
5. Keith Henney, "Radio Engineering Handbook", McGraw-Hill, 1959, pp. 14-42 – 14-43.

**ham radio**

# further automation for typewriter-type electronic keys

Robert L. Kurtz, W6PRO, Micro-Z Company, Box 2426, Rolling Hills, California 90274

Some time ago I developed an automatic typewriter-type keyer that features self-completing characters, automatic character spacing, wide speed range and includes special characters used by hams.* In response to several queries the possibility of adding another bank of memory storage within the same keyer case was investigated; this would permit the operator to type ahead of the actual transmission—or to slow down momentarily while hunting for a key—and still produce faultless Morse code.

With an added "buffer storage module" you type at the same average speed as the keyer (as normal), but if you hit two keys close together and a third much later, the keyer sends perfect code. The buffer storage results in several subtle changes. In the basic keyer, for example, a key is held down to repeat a character. With the buffer storage, you tap the key twice just as when using a typewriter to repeat a character. If you press a key before the logic circuits are ready for it, with or without the storage module, you cannot affect the stored characters; simply hold the key down, and when the memory circuit is clear, the new character is stored automatically, and the key can be released.

With the added buffer storage, two keys can be tapped rapidly and a third key held down until the first character is complete, a fourth key held down until the second character is complete, etc. Your average typing speed is not increased, but you have a backlog of two characters that gives you a much wider operational flexibility.

## basic keyer

The basic keyer consists of two circuit boards (matrix and logic board) and a keyboard. The matrix board contains the information on each character and mounts on the keyboard. The logic board stores the information on the selected character, scans this storage at the selected speed, provides the proper spacing between characters and contains the output relay and side-tone circuitry. A simplified block diagram of the basic keyer is shown in fig. 1.

Thirteen wires carry the information from the matrix board to the logic board - one trigger wire to signal the logic board that a key has been pressed and six pairs of dot/dash wires that carry character information. This permits the formation of

*This keyer is marketed as the Pro-Key Kit and is available from Micro-Z Company, Box 2426 Rolling Hills, California 90274 for $149.50 postpaid. The buffer storage unit described in the article is available as a kit for $34.50 additional.

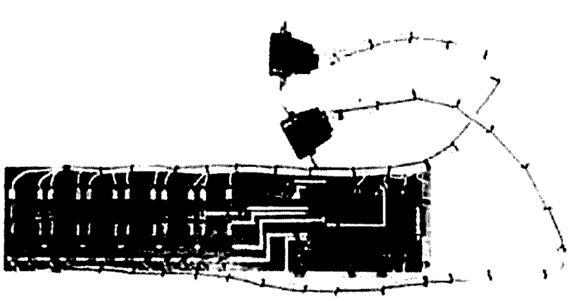
characters with up to six dot/dash combinations such as SK, period and question mark. If the "A" button is pressed, a positive voltage appears on the "dot" line of the first pair of wires, on the "dash" line of the second pair of wires, and on the "trigger" wire.

The trigger signal starts the keyer clock and opens the insert gates momentarily to store the selected character in the main storage elements on the logic board. After the gates close, the stored character cannot be disturbed. The logic circuits then scan the main storage to create the Morse character. If a new key is held down, the new character is automatically inserted only after the first character is complete and the proper spacing time has elapsed.

## buffer storage module

The additional storage module is inserted between the keyboard and the main logic board (fig. 2). When a key is pressed, the trigger signal is differentiated, stored in the module, and is used to open the module signal input gates so the selected character can enter the buffer memory.

As soon as the selected character has been entered into secondary storage, the trigger memory is erased, and the input gates close. The presence of a stored character in the module generates a trigger signal which is sent to the basic keyer. This starts the clock and inserts the character in the main storage, just as if the signals were coming directly from the keyboard. The pulse that opens and closes the main insert gates on the basic keyer is used to clear the secondary storage in the storage module (after a slight delay). The total process takes only a few microseconds, and the storage module is ready to accept another character from the keyboard. Meanwhile, the logic in the basic keyer is leisurely scanning the main storage to create the first character at the selected sending speed.



**Buffer storage module**

Another key can be tapped immediately and the buffer storage process is repeated. However, since the first character is still being transmitted, the second character remains in storage until the main storage in the basic keyer is ready to accept it. The presence of a stored signal in buffer storage "locks out" — or inhibits — the storage module input gates so the second character cannot be disturbed until it moves into main storage and the buffer storage is cleared.

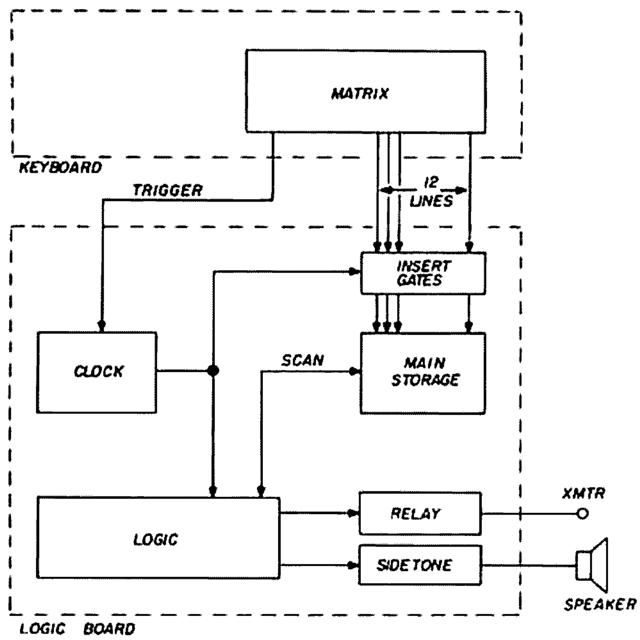



**fig. 1. Block diagram of the basic pro-key typewriter-type electronic keyer.**

A third key can be pressed immediately after the second. When the keyer has completed the first character, the character in buffer storage automatically moves into main storage, and the buffer is cleared. The buffer storage will then accept the third character, the third key can be released, and the operator can move on to the fourth character while

the second is being transmitted.

An entire CQ or DE is stored by simply tapping two keys, just as you would on a typewriter, and the Morse characters are des are raised above ground. This is accomplished by a positive trigger signal from the keyboard operating through the trigger storage flip-flop and inhibit

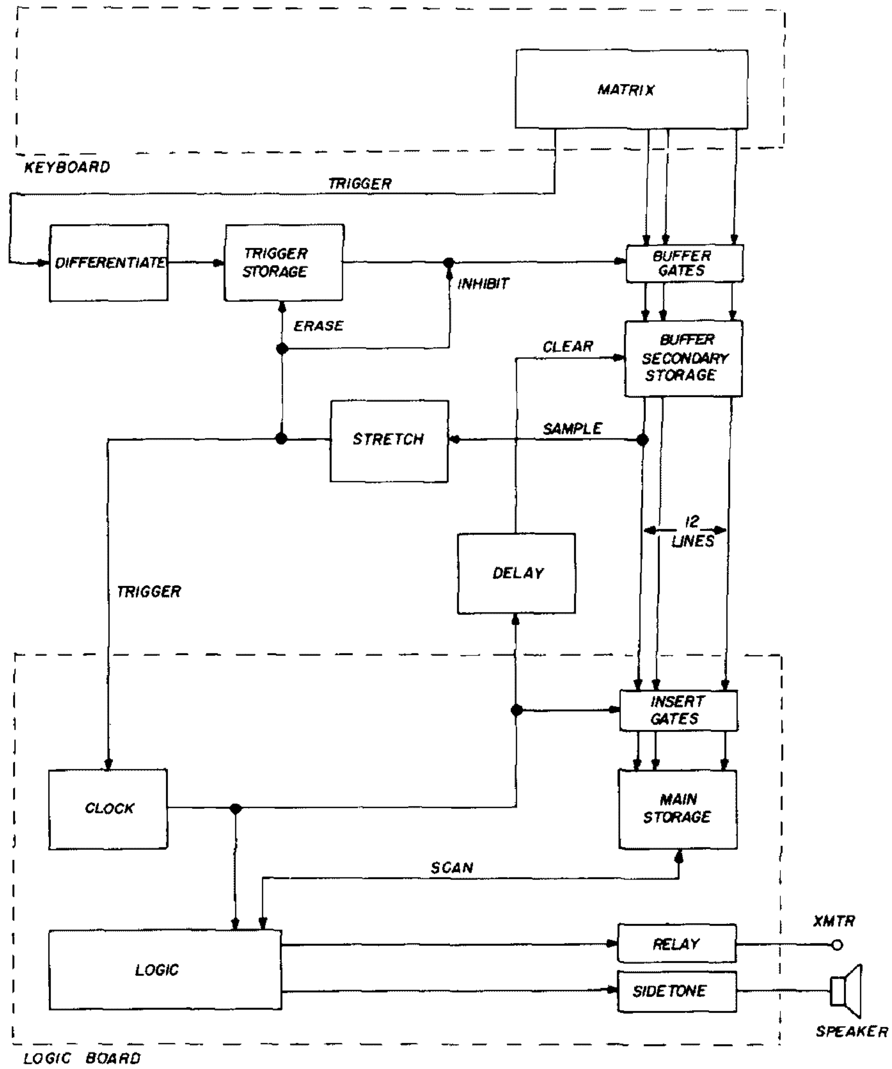



**fig. 2. Block diagram of the keyer with the added buffer storage module.**

transmitted at the speed set by the operator. The operator can speed up or slow down his typing but the keyer transmits smooth, even code. This is particularly useful for hunt-and-peck typists who use a typewriter-type keyer at 20 to 25 wpm.

## circuit

In the logic diagram of the buffer shown in **fig. 3**, only three of the twelve storage RS flip-flops are shown for simplicity. Keyboard signals set the proper storage flip-flops to provide a positive output only if the cathode ends of the gate diodes are raised above ground.

gate to apply a positive signal to all gate diode cathodes. The first two storage flip-flops are sampled by a two-input NOR gate which resets the trigger flip-flop and closes, or grounds, the input gate diodes through the inhibit gate when either of the two storage flip-flops have been set. The output of this NOR gate is stretched by the series 5k resistor and .01 $\mu$F capacitor to give all of the appropriate storage flip-flops a chance to set before closing the input gates.

If another key is pressed while the buffer storage is full, the trigger flip-flop is

again "set" and tries to open the input gates, but the inhibit gate prevents this until the storage flip-flops are cleared by the main logic board. The NOR gate that fer. This provides the delay necessary to assure sufficient time to insert the stored signal in the keyer prior to clearing the buffer.

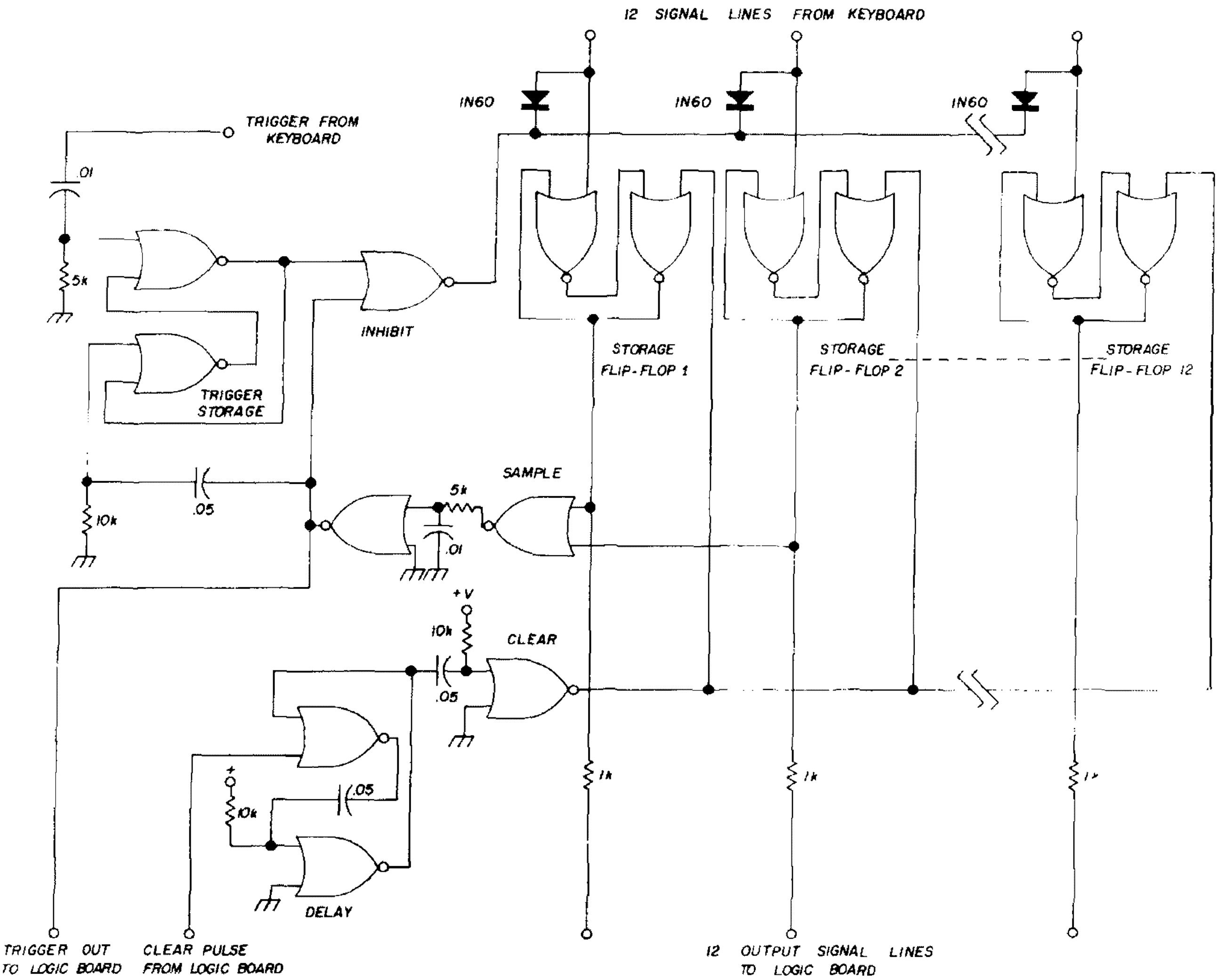



**fig. 3. Logic diagram of the buffer storage unit. The author used eight Motorola MC724P quad two-input gates in his model.**

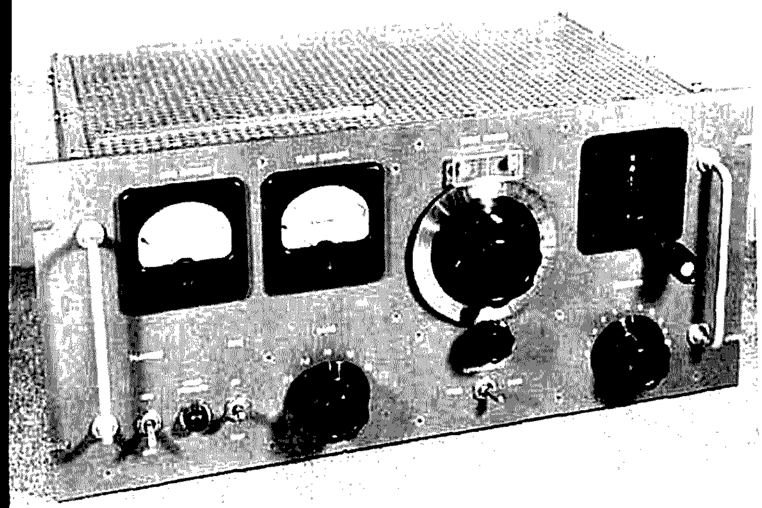
detects the presence of a stored signal and inhibits the opening of the input gates also supplies a positive trigger to the basic keyer, just as if the signals in buffer storage were coming directly from the keyboard. With this logic, if a key is pressed while the storage is empty, the character will be inserted only once, and the key must be pressed again for a repeat.

When the main insert gates are opened, a positive pulse triggers a one-shot circuit. The negative-going trailing edge of the output pulse of this stage is inverted and used to reset all flip-flops in the buffer.

## adding more storage

Since the output of the buffer storage appears as a virtual keyboard, and is compatible to input and output signal logic polarity, buffer modules can be stacked to provide banks of four, five, six or more character storage. I have connected four of these together to store any five-letter word (such as my call) but the utility is questionable since the first character is generally complete by the time you are typing the third or fourth character, except at very low code speeds.

ham radio



photographs by Ted Stites

# homebrew five-band linear amplifier

A conservatively designed circuit using time-proven 811-A's

Harry R. Hyder, W7IV

It is customary to preface a construction article with a few remarks about why the author decided to build rather than buy the equipment described. In my case, there's only one reason why I build radio equipment: I enjoy it.

I don't enjoy hole drilling or coil winding any more than an artist enjoys mixing paint or cleaning brushes. My satisfaction comes from creating something unique from my own mind and hands.

I read the construction articles in ham radio and other magazines every month, but I've never built equipment that exactly duplicates a published description. What I look for is not something to copy, but rather the construction hints and ideas that I can adapt to my own requirements.

This article is presented in that spirit. You may not wish to copy this linear amplifier, but you could do worse. Perhaps you'll find something you can use in your next construction project.

## circuit description

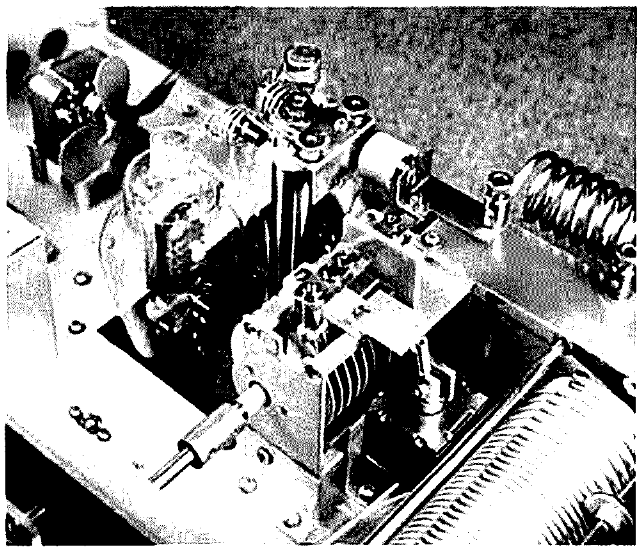
Parallel 811-A's are used in a grounded-grid circuit (fig. 1). In terms of watts-per-dollar of tube cost, the 811-A must head the list. Some hams complain of a short life for these tubes when operated at ICAS ratings as these are; however, I find it's easier to buy a couple of inexpensive tubes frequently rather than a single expensive tube occasionally.

The cathode circuit has a matching network to transform the 50-ohm input to approximately 150 ohms required by the tubes. A cathode matching network is often dispensed with, but it has its virtues. A 3:1 mismatch is frequently beyond the capability of some exciters. If the exciter doesn't have some power to spare, it may not be possible to drive the amplifier to full output without the network. With the matching network, the transmission line is "cold" and may be of any reasonable length. Some writers have reported that the matching network also improves amplifier linearity. Therefore, since it's simple and requires no tuning, it's cheap insurance.

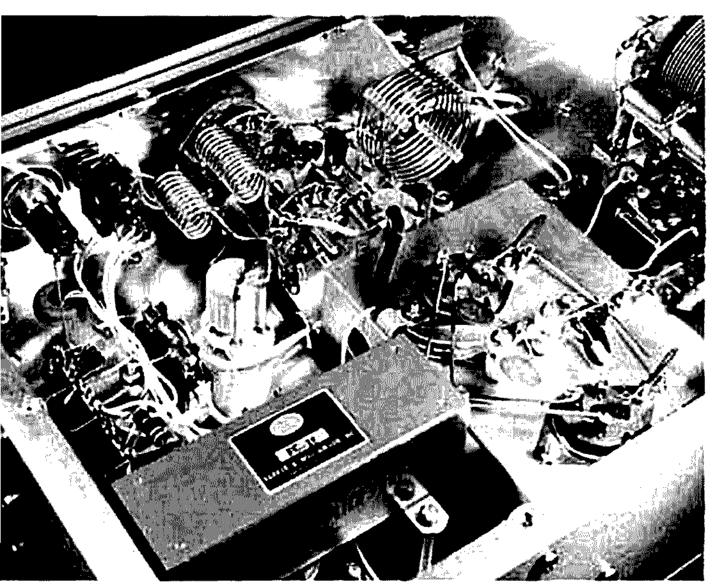
The network is an L configuration on 80, 40 and 20 meters, changing to a pi network on 10 and 15 meters. The high effective cathode-to-ground capacitance, consisting of tube and wiring capacitance plus the distributed capacitance of the filament choke, precludes the use of an L network on the two higher frequency bands. The tapped 20-, 40- and 80-meter cathode inductance is in the circuit at all times. On 10 and 15, small self-supporting airwound coils are connected in parallel with it. This is merely a switching convenience.

The plate tank coil is a roller-type inductor for the low-frequency bands, with a series-connected small coil for 10 meters. The variable inductor permits adjustment for optimum Q on all frequencies.

Detail of the amplifier tank circuit. The small coil in the binding posts is the 10-meter inductor.



Circuit details and component layout of input section. Attention to detail results in a professional appearance.



The plate tank capacitor is from a BC-375 tuning unit. Its original capacitance range was 23 to 140 pF. I wanted to reduce minimum tank capacitance on the high-frequency bands to lower the loaded Q and increase efficiency. I carefully split the stator with a fine saw. Only one of the sections is used on the high-frequency bands, reducing the minimum tank capacitance by about 12 pF. This decreases the loaded Q on 10 meters from 26 to 20, and on 15 meters from 19 to 15. The photos show the switching arrangements to cut in the second section. The contacts are from an old relay, and the solenoid is a 115 Vac unit I happened

to have in my junk box. The solenoid is controlled by a front-panel switch.

The loading capacitor is a five-gang 420-pF-per-section unit that came from an MN-26 radio compass. Two sections in parallel are used on the higher frequencies; the remaining three are cut in by a relay controlled by the tank capacitor switch. The capacitor is available from Barry Electronics.

At 1500 volts, 811-A's require about 4.5 volts bias, which is supplied by a 4.7-volt zener in the filament return. This is less expensive and more reliable than a bias supply, and has a very low impedance. A 100-volt zener is also in the filament return, with a small amount of dc current bled through it. This provides full cutoff bias. It can be cut out by a front panel switch, or by external relay contacts.

The plate-current meter is also in the filament return, but reads plate current only; not total cathode current. The grid-current meter is in the dc grid return.

The high-voltage bleeder consists of four 150-k ohm 2-watt resistors in series, since it is not good practice to put more than about 500 volts across a single 2-watt resistor. I like redundant bleeders; should the one in the power supply open, the one in the amplifier will discharge the filter capacitors in a few seconds. A neon lamp indicates high voltage on the amplifier.

## construction

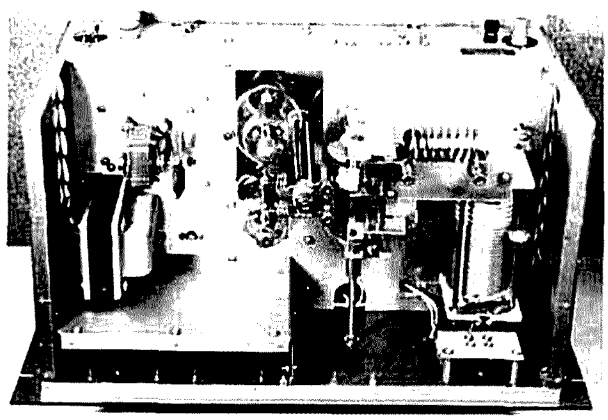
The chassis is aluminum, 10 x 17 x 3 inches. The 811-A's are mounted on a 4 x 6 x 1½-inch aluminum chassis upside down. I made these chassis sides and the meter shields from pieces bought in a scrap-metal yard.

The cover shield is cane-pattern sheet aluminum from a "do-it-yourself" department of a hardware store. This material is rather flimsy, so I stiffened it and improved the rf shielding with ½ x 1/16-inch aluminum strips on the outside. The ½ x ½ x 1/16-inch aluminum angle stock that holds the shield assembly was also obtained in the scrap-metal yard, but the same material is sold as trim in most hardware stores.

## wiring

All power and control wiring should be installed first. Plan the wiring so that when the individual wires are joined *into* cables, the cables will run parallel to the main chassis dimensions. Strip each wire and tin it at both ends before placing it into the chassis. Leave a generous "service loop" when determining length; this makes parts replacement easy.

Lacing the cables adds a lot to appearance. Flat nylon ties are good. Start at the cable center and work toward the ends, bringing out individual wires as required.

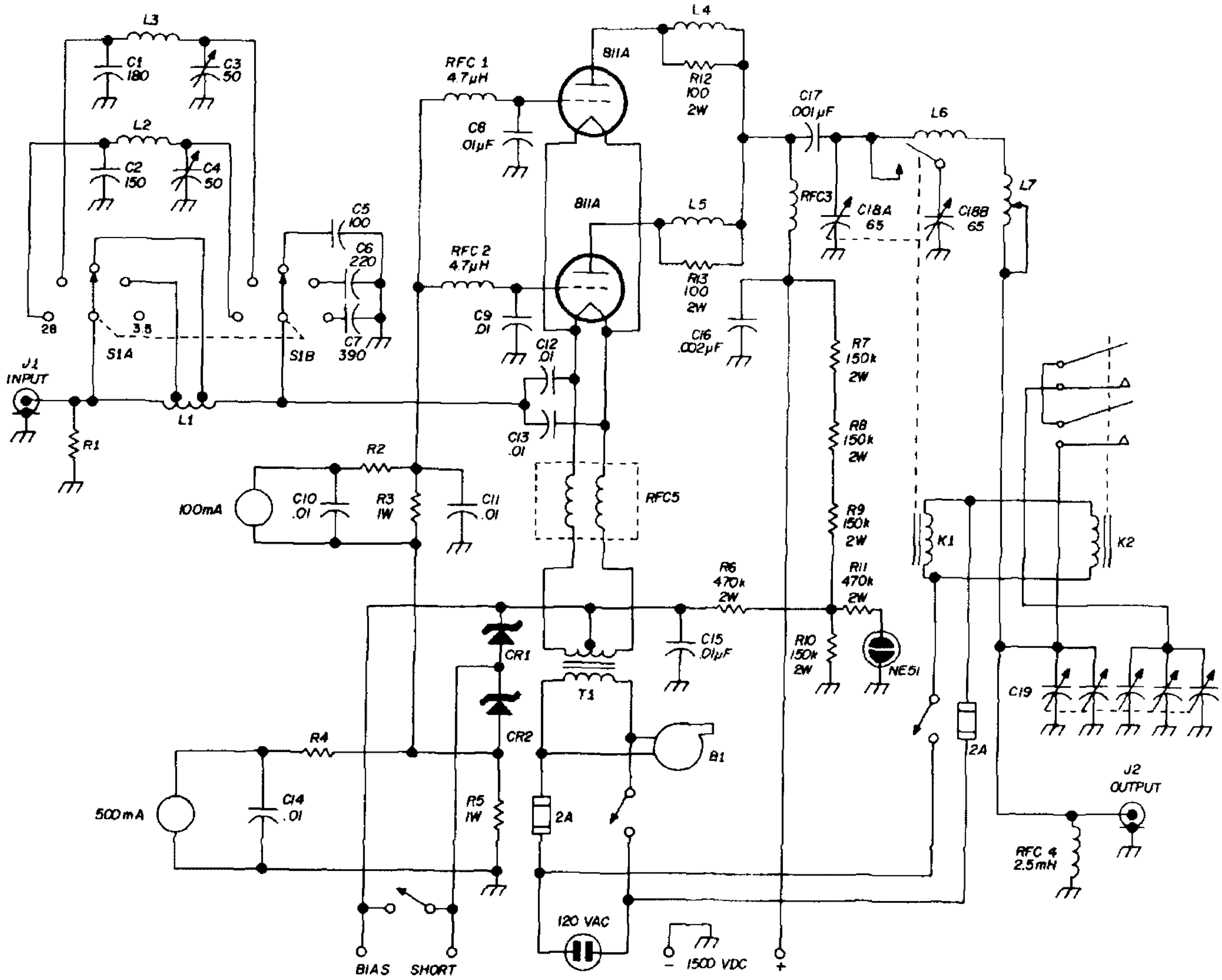


Conductors in low-level rf circuits consist of bare tinned bus bar. Output circuits are brass or copper strip about 0.02-inch thick. These strips should be secured with screws and nuts rather than solder. For appearance, sand the strips and spray them with clear lacquer.

## the panel

I prefer gray wrinkle to all other finishes. I purchase a blank panel with a black-wrinkle finish, complete all drilling, then spray it with "machine gray" lacquer. Several light coats are better than one heavy coat; the lacquer adheres better, and there's less tendency for the lacquer to fill in the original black finish. This makes for color standardization, because no two gray-wrinkle panels are of the same hue, even from the same manufacturer's lot.

Another finish, used on my amplifier,




| | |
|---|---|
| B1 | Cooling fan (Japanese import; see photo) |
| C18A,B | Variable, 2 section, 65 pF per section, 0.07 = inch spacing |
| C19 | 5 section, 420 pF per section |
| K1 | See text |
| K2 | Relay, dpst, 10 A contacts, 117 Vac coil |
| L1 | 7 = ½ turns 1 = ½ inch diameter, 2 inches long, tapped 3rd and 5th turns. Approximately 4.5 uH total inductance, tapped at 2.4 uH and 1.2 uH |
| L2 | 9 turns number 14, 5/8 inch ID, approximately o.8 uH |
| L3 | 12 turns number 14, 5/8 inch ID, approximately 1.0 uH |
| L4,L5 | 3 turns number 14, 5/8 inch ID, wound around R12 and R13 (see photo) |
| L6 | 8 turns 1/8 inch copper tubing, 3/4 inch ID, 2 inches long |
| L7 | Inductor, variable, 18 uH maximum (E.F. Johnson 229-202) |
| R2,R4 | Adjust for correct reading of M1 and M2 |
| RFC1,RFC2 | 4.7 uH pigtail |
| RFC3 | 90 uH, 500 mA (B & W |
| RFC4 | 2.5 mH, pie wound |
| RFC5 | Filament choke (B & W FC-15) |
| SW1 | 2-gang rotary, 2 poles, 5 position |
| SW2,SW4 | Spst toggle switch |
| T1 | Filament transformer, 117 V pri., 6.3 V 10 A sec., CT (Triad F-21A) |

**fig. 1. Schematic of the 811-A grounded-grid linear amplifier. Matching section in cathode circuit provides a 3:1 transformation ratio, assuring adequate drive from most exciters.**

requires nothing but a wire brush. Clamp the piece to a flat surface and make straight, even strokes with the brush. It produces a beautiful grained finish.

Whatever finish you use, handle the pieces with cloth gloves — fingerprints

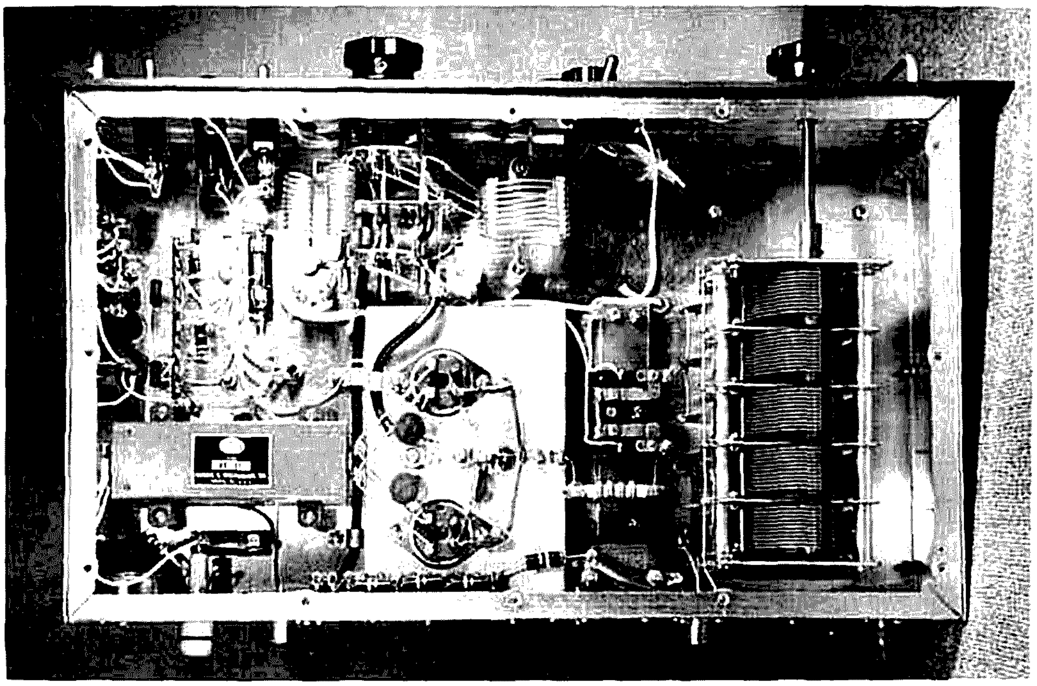
really stand out. Dust off the pieces and give them a couple of light coats of clear lacquer. Surfaces to be joined should be masked to obtain good electrical contact.

## accessories

The fluted knobs and nickel-silver dial may look old fashioned, but I like them. They're still available commercially. The dial pointer was lost years ago, so I made one from a scrap of plastic. The pinch drive provides just enough drag to keep the [illegible] capacitor from getting out of adjustment.

The meters are su[illegible]ms. Their sensitivity wasn't what I wanted, [illegible] this was corrected using standard techniques.[1]

The roller-coil dial is homemade. I bought a 3-digit counter from a surplus dealer for a dollar. The miter gears were obtained from a standard right-angle drive. I cut the escutcheon from 1.8-inch-thick sheet aluminum. It's finished in black-wrinkle lacquer. A possible source of wrinkle finishes in spray cans is your neighborhood Speed Shop; the hot-rod set seems to favor these finishes nowadays.



Bottom view of the linear amplifier. Note lead dress and method of securing cables.

## decals

You'll want to label your controls and other accessories. I prefer the water-type decals to the dry-transfer labels because mistakes are easier to correct. With the latter, you're committed to a position on the panel, and it's difficult to remove dry-transfers without ruining the finish. After you've positioned the decals, spray them with clear lacquer.

## a final word

If this is one of your first major construction projects, and you've made a few mistakes in mechanical work, all is not lost. Most goofs can be remedied. Extra holes can be occupied with screws and solder lugs, as if this is what you intended all along. Or you can strip the finish and fill the hole with auto-body solder, then refinish the panel. This takes a few hours of extra work, but it reflects your pride in a job well done.

## references

1. "The Radio Amateur's Handbook," 46th edition, 1969, American Radio Relay League, p. 528.

ham radio

# a new approach to equipment rack construction

Here's a low-cost rack that can be easily built to accept any panel width

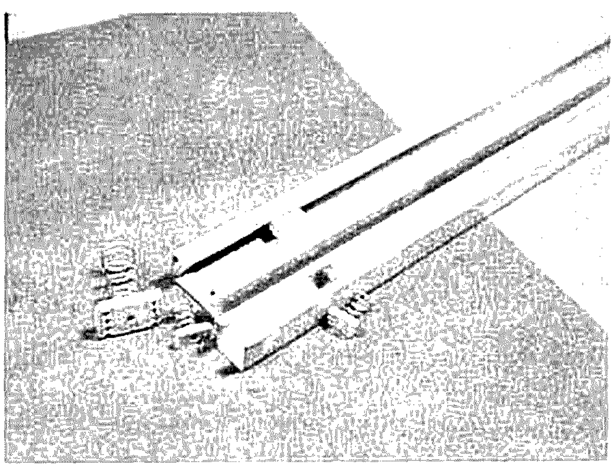
George R. Allen, K1EUJ, Route 2 Robin Circle, Tolland, Connecticut 02891

Equipment racks are a useful asset to almost any amateur station. Known also as "relay racks," they are available commercially only in sizes for standard 19-inch panels. This article describes a simple, inexpensive approach to rack construction whereby racks can be built to accept panels of any size. Material costs range from $4 to $8 per rack.

## design

Commercially available racks consist of a metal frame with the front portion drilled and tapped to accept 10/32 screws

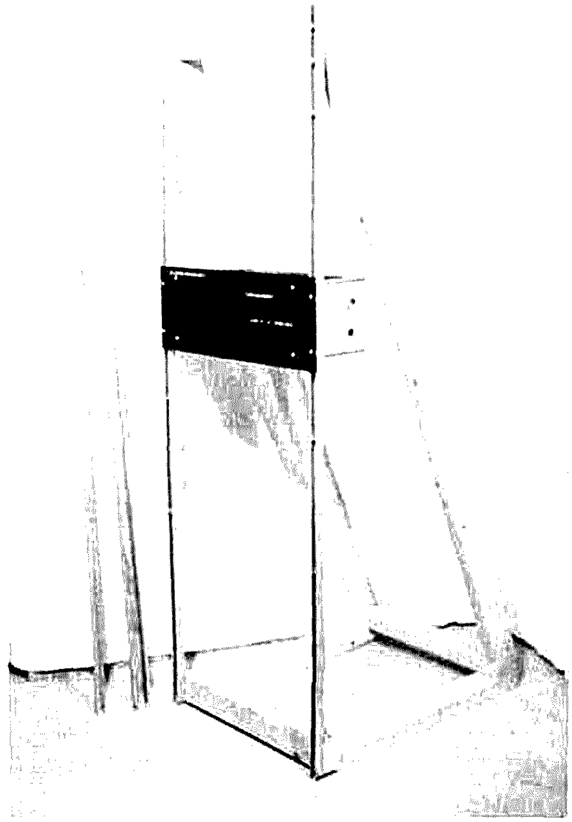
Two sizes of steel channel with spring-loaded nuts for mounting panels.



for panel mounting. The racks I've built consist of a wooden frame with two pieces of channel steel attached to the front for panel mounting. The channel steel, called *WHIZ STRUT*, comes complete with spring-loaded nuts that may be inserted into the channel at any location where a panel is to be attached to the frame. The spring, attached to the nut, forces the nut against a flange on the channel, causing the nut to stay in place. The photo shows large and small sizes of strut with associated spring-loaded nuts. These nuts are for 10/32 screws; however, other sizes are available.

## a simple homemade rack

A sketch with dimensions for constructing the rack shown in the photo is given in fig. 1. This rack is 58 inches high and accepts 19-inch panels. However, there's no reason why the width can't be



Homemade equipment rack with 19-inch panel spacing. Panels of any size can be accommodated by varying width of channel supports.

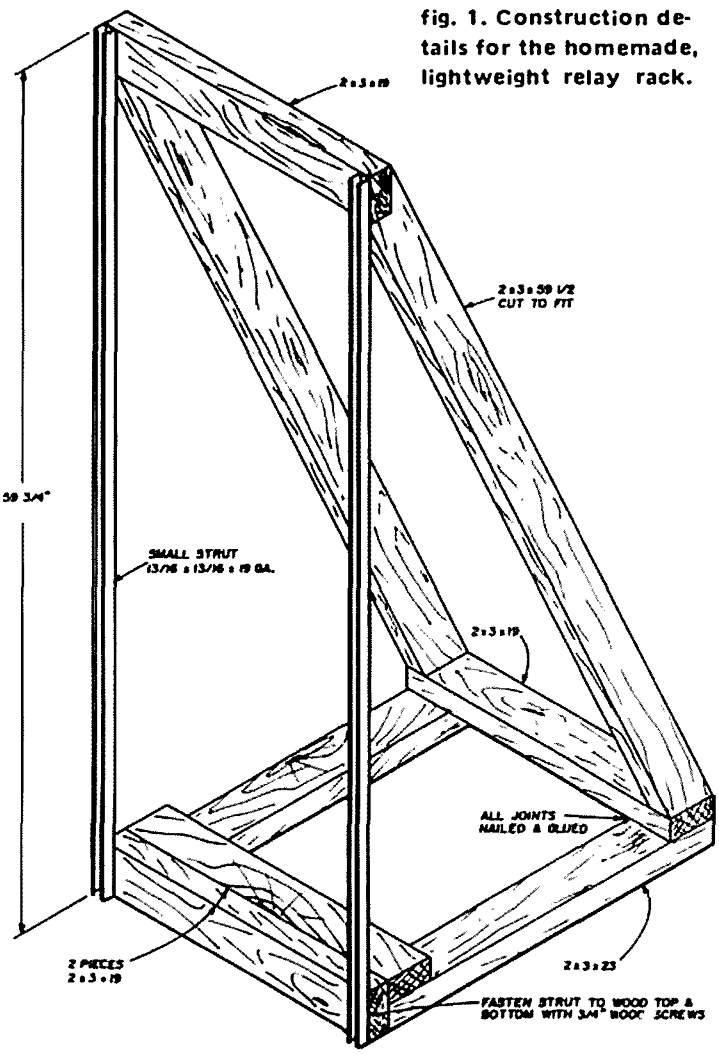



fig. 1. Construction details for the homemade, lightweight relay rack.

changed to accept panels longer or shorter than this. The cost of the rack in the illustrations was $5 for the struts and spring-loaded nuts, and about $2 for the lumber.

Heavy equipment can be mounted easily in this rack by one person. The equipment is loosely fastened to the nuts while the equipment rests on the bottom of the rack. The equipment is then slid up the rack until it's in the proper position, where it will stay with little effort. Once in position, the equipment may be securely fastened.

Struts and hardware may be obtained from the distributor, who will send you a price list upon request.*

*Donald S. Tunnel, Box 331, Fort Washington, Pennsylvania 19034. Ask for amateur price sheet "A".

ham radio

# solid-state radio direction finder

Sam Kelly, W6JTT, 12811 Owen Street, Garden Grove, California 92641

Radio direction finding has come a long way since F. Braun did his "Research on a Method of Directive Wireless" early in 1903. One of several early experimenters, Braun conducted his DF work with the German army balloon department.

Much early DF work was confined to observing weather phenomena and experimenting with antenna directivity. The Russian scientist Popov, for example, was primarily interested in tracking thunderstorms. By 1911, Italians Bellini and Tosi were experimenting with DF on shipboard. Military requirements spurred DF development during World War I.

From 1918 until the late 1940's, radio direction finding advanced with improvements in receivers, antennas, and servo systems. Today, commercial DF is giving way to more advanced navigational aids such as satellite repeaters, Loran-C, Omega, Decca and inertial systems. Radio direction finding, however, is still important for navigation in small vessels and as a backup where long-range navigation systems don't provide coverage.

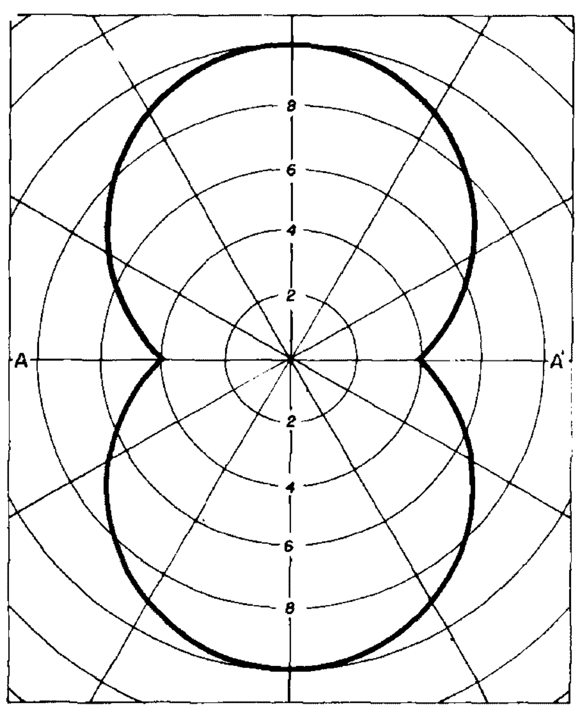
This article describes a small DF adapter for communication receivers, using a simple diode detector and transistor audio amplifier. It's useful for locating radio interference sources, hidden transmitters in radio club contests, or as an addition to your boat's communication equipment.

## df principles

Radio direction finding is based on the directional characteristics of antennas. The idealized pattern of a loop antenna is shown in **fig. 1.** Under ideal conditions, this pattern can be plotted by measuring the voltage at the antenna's terminals as it is rotated 360 degrees in the horizontal plane. A maximum voltage is obtained at the antenna terminals when the transmitter is in line with the loop.

The minimum voltage, or null, is obtained when the loop is broadside to the transmitter. The nulls are of great importance, as they provide the most sensi-

**fig. 1. Idealized pattern of loop antenna. Line A - A' is in direction of transmitting station.**

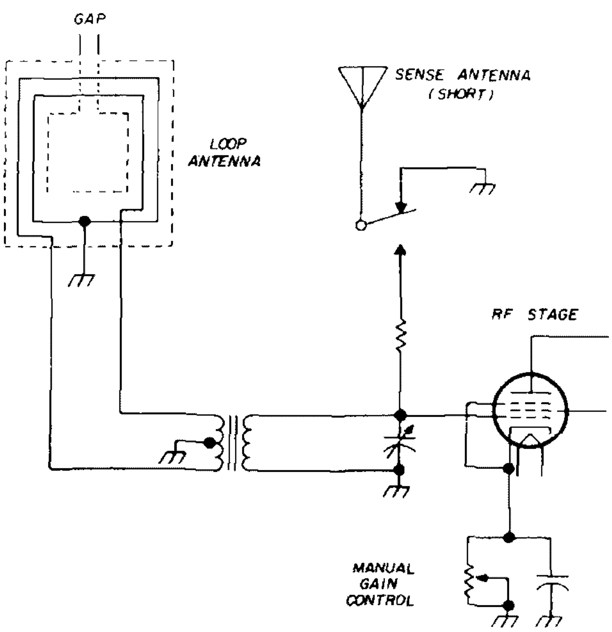


tive means of determining the signal path.

You can see from **fig. 1** that a peak or null can be obtained with a transmitter in either of two directions with respect to the loop. This is called the "180-degree ambiguity."

In the early days of radio direction finding, the 180-degree ambiguity caused many accidents. A conventional DF antenna is shown in **fig. 2.** A separate vertical or "sense" antenna is used with the loop to resolve the 180-degree am-

**fig. 2. Input circuit of direction finder. Sense antenna modifies loop pattern to that shown in fig. 3.**




biguity of the transmitter location. In practice, the operator adjusts the loop for minimum signal with the sense antenna disconnected. He then finds the sense, or general direction, of the transmitted signal by rotating the loop 90 degrees with the vertical antenna connected. If the addition of the vertical increases the signal in one direction, then minimum signal indicates that the transmitter is in the opposite direction. Adding the vertical antenna changes the pattern of the basic loop to a cardioid(**fig. 3**).

The cardioid pattern is much too broad to give an accurate bearing. It is used only to sense the general direction of the transmitter.

## the goniometer

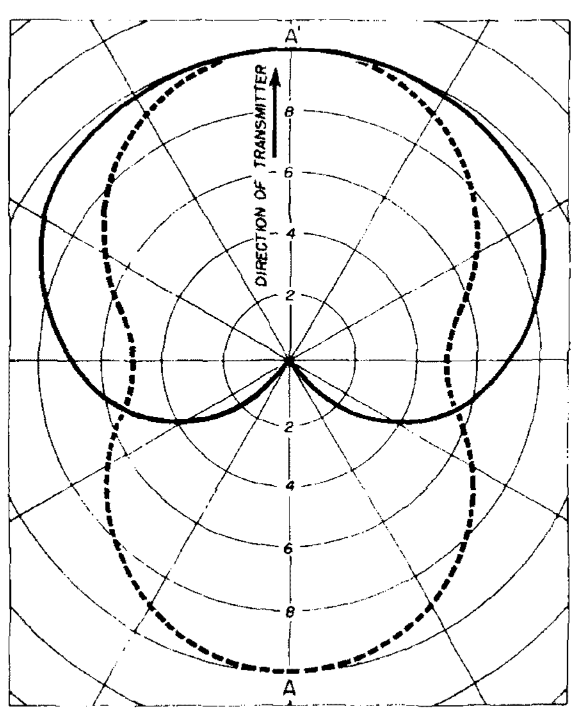
An interesting variation of the DF loop is the Bellini-Tosi antenna shown in **fig. 4.** This consists of two fixed loops at right angles to each other. Loop outputs are coupled to two identical rf coils accurately mounted at right angles to each other. A secondary coil is free to rotate around the fixed coils, making a rotary transformer to couple the output to the receiver. The shaft of the secondary is attached to a calibrated dial.[1]

The Bellini-Tosi system has several advantages such as convenient shipboard mounting (all rotating components can be below deck), and the ability to match a long transmission line. It can also be used at fixed locations, since the torque required to turn the shaft of the rotary transformer is within the capability of typical synchros.

## df adapter

The ferrite loops used in transistor radios can be readily modified for DF work. A small DF adapter for use with a

**fig. 3. Cardioid pattern of loop caused by addition of vertical sense antenna. Loop is rotated 90 degrees to determine "sense", or general location, of transmitted signal.**



communications receiver is shown in the photo. Its schematic is shown in fig. 5. The crystal detector and transistor audio amplifier allow the adapter to be used without a receiver. It can be used to locate sources of interference or hidden transmitters in club activities. The signal must be either broadband noise or an a-m transmitter, since there's no provision for product detection.

using a signal generator. The sense antenna is a small transistor-radio whip. The loading coil was salvaged from a DAV antenna. The loading coil is peaked on the frequency to be tracked by tuning in the signal, depressing the sense button, and adjusting the tuning slug for maximum signal.

Inexpensive British surplus headphones were used. These have a very low

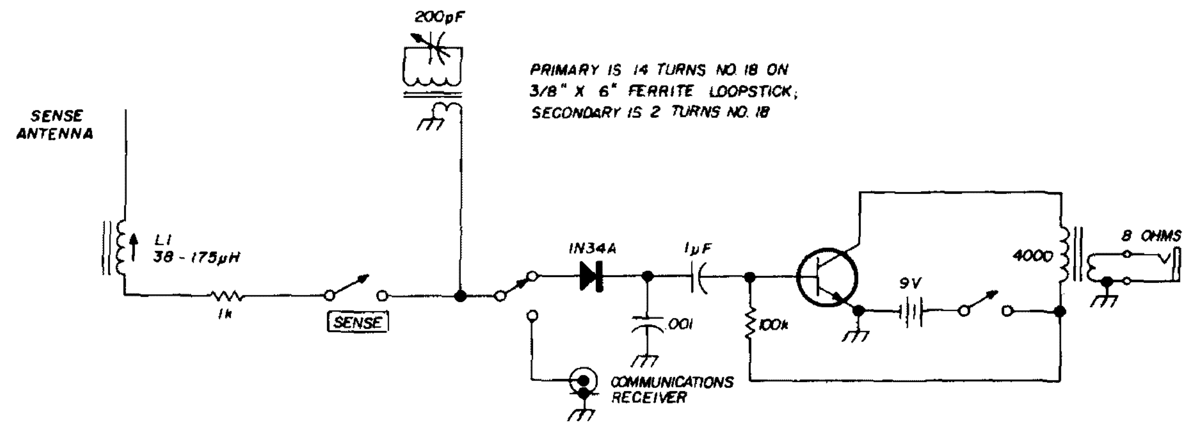



fig. 5. Direction finding adapter schematic. Sense antenna is a piece of vertical wire or small tubing 3 - 4 feet long; loop is made from transistor radio ferrite loopstick.

## construction

The adapter was built in a small aluminum box. Care was taken to shield the sense lead and the lead from the switch to the coaxial connector for the external receiver. The dial was calibrated

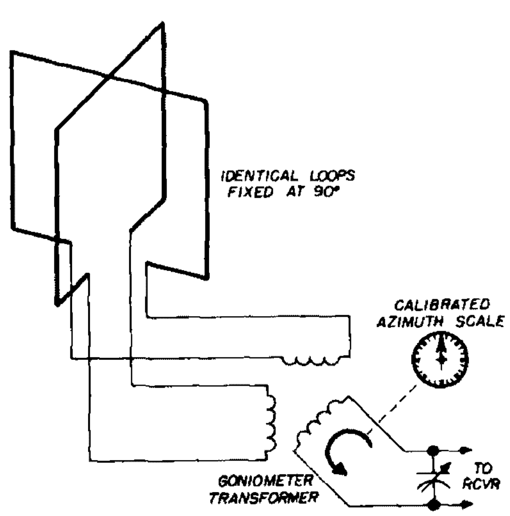


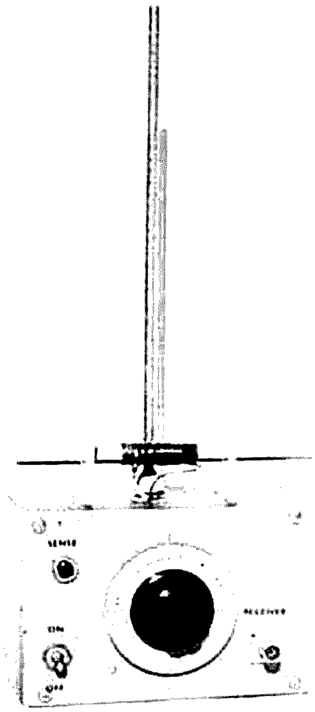

fig. 4. System using the Bellini-Tosi goniometer principle. Loops are fixed; goniometer transformer is rotated to obtain bearing.

impedance so the transformer is required. If you are using 2000- to 4000-ohm magnetic headphones, the transformer can be omitted.

## operation

As you gain experience, you'll quickly learn the limitations of direction finders. There are two major problem areas: location problems and multipath signals due to propagation phenomena. Take bearings from high points in flat, unobstructed areas. Remember that bearings taken in or near buildings having a lot of metal will be inaccurate. Operation in dense foliage or rough terrain will also cause errors. The easiest way to determine if the bearing is wrong is to take bearings at frequent intervals. Then any odd bearing will be obvious.

Night effect is quite pronounced in the 2- to 8-MHz region. This is a propagation phenomena resulting in changes in the polarization of the sky wave signal. The symptoms can range from complete



DF adapter for use with communications receiver or as a "walk-around" sensor.

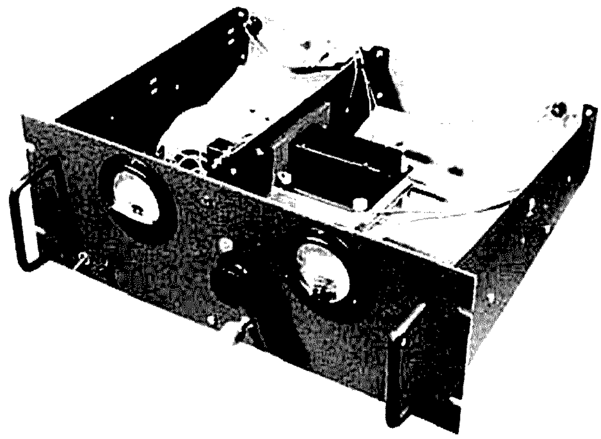
absence of a null to pronounced nulls as much as 90 degrees from the correct bearing. This effect is most pronounced at sunrise and sunset – periods that correspond to the most rapid changes in the height of the ionosphere. Fortunately, the effects of the polarization changes are reduced the closer you are to the transmitter. In most transmitter hunts these aren't evident. If you have plenty of space, you can reduce the night effect by using a vertical antenna system such as the Adcock.

Like most crafts, the best way to become proficient in direction finding Is to practice. How long has it been since your club had a hidden-transmitter hunt or tracked an irritating source of radio frequency interference?

### reference

1. F. E. Terman, "Radio Engineers' Handbook," McGraw-Hill, New York.

ham radio



# a power amplifier for 1296 MHz

Dual planar triodes in a half-wave resonant cavity provide 100watts output with 10 dB power gain

R. E. Fisher, W2CQH, C. W. Schaible, W2CCY, G. W. Schober, W2OJ, R. H. Turrin, W2IMU

One reason for the limited amount of activity on the uhf bands is lack of commercially available equipment or proven designs for homebuilt projects. This article presents a grounded-grid power amplifier for 1296 MHz using two 3CX100 A5/7289 planar triodes. Easy-to-work materials are featured, and very little soldering is required. Although material substitutions and modifications to the construction methods are feasible, we recommend that the amplifier be built according to the directions given. This will ensure correct performance and a minimum amount of debugging.

The amplifier will deliver a minimum of 100 watts into a 50-ohm load, at 50 percent efficiency, with a power gain of 10 dB. The circuit is similar to one described in reference 1. The difference is mainly in the placement of tubes in the cavity and in the cavity tuning circuit. The initial design, developed by W2CCY, was water cooled and very rugged. A later version, built by W2CQH, featured air cooling and simplified construction. Tests have shown that air cooling is adequate, but quieter operation is obtained with water cooling.

## circuit description

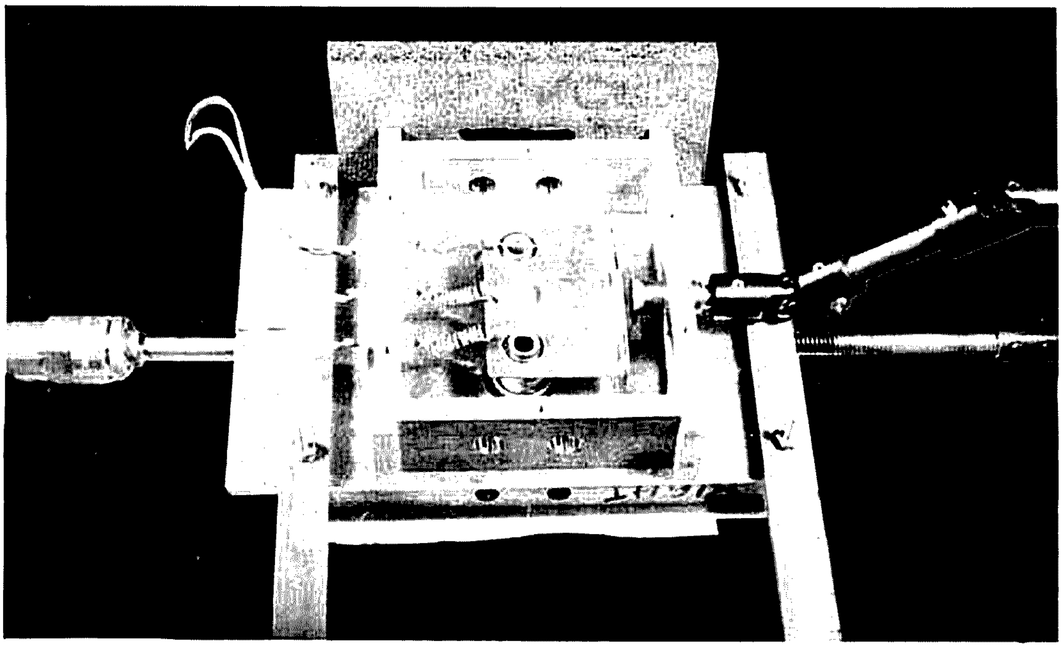
A conventional grounded-grid circuit is used, with bias provided by a zener diode in the cathode (fig. 1). A 27-volt, 10-watt zener is satisfactory for class-C amplifier operation. However, partial or full resistor bias, which has the advantage of being fully adjustable could be used. The resistance must be bypassed with a large capacitor for a-m operation. This arrangement isn't recommended, because each time the amplifier is turned on it receives no bias initially. Thus, it will draw heavy anode current until the capacitor charges.

The zener, on the other hand, provides immediate bias and also presents a low impedance at audio frequencies. Since grid current flows through the cathode-bias circuit, a zener minimizes drive power loss that would occur across a cathode resistor. Several zeners may be connected in series if a single unit isn't available.

## anode cavity

The anode cavity is a loaded half-wave resonant circuit, which extends along the centerline through the tubes. Immovable brass bars, which form the sides of the cavity, determine the approximate resonant frequency. A sliding plunger allows vernier tuning near 1296 MHz.

A magnetically coupled link provides a wide range of output loading. The input circuit uses a low-inductance strap that ties the cathodes in parallel and extends as a short stripline, which resonates as a half-wave tank. Approximately one-quarter wavelength of this tank is the internal cathode structure of the tubes. Capacitive tuning is provided at the other end of the tank. Input coupling is by means of a capacitive probe. Since the external connections to cathodes and heaters are in a low-rf field, simple wirewound rf chokes can be used in the bias and heater lines.



**Inside view of cathode cavity showing the panel-bushing tuning capacitor. Sponge rubber plenum is visible at the top.**

Metering is included in both anode and cathode leads to maintain the grid at earth potential. Both meters are necessary to determine grid current. Grid current is simply the difference between cathode and anode current. Grid current is a sensitive indicator of correct operation when adjusting tuning, loading, and drive for optimum efficiency.

## construction

Both anode and cathode resonant cir-

cuits are constructed of brass sheet and rod, which are available at most metal supply houses and some hobby stores. The brass rod used for cavity walls has uniform dimensions and a smooth finish, therefore no soldering is required at the joints. Instead, 4-40 or 6-32 machine screws on 3/4-inch centers provide sufficient contact.

Both the plate cavity top cover and by-

Fig. 2 shows the plate cavity interior and grid contacts, which are soldered in finger stock similar to the plate contact arrangement. The sidewalls of this cavity are constructed of 1/4-inch thick brass rods, which are 5/8-inch high. The tuning plunger is made of 1/4 x 1/2-inch brass rod, with finger stock soldered to the top and bottom edges. The plunger assembly is moved by a 1/4-inch-diameter threaded

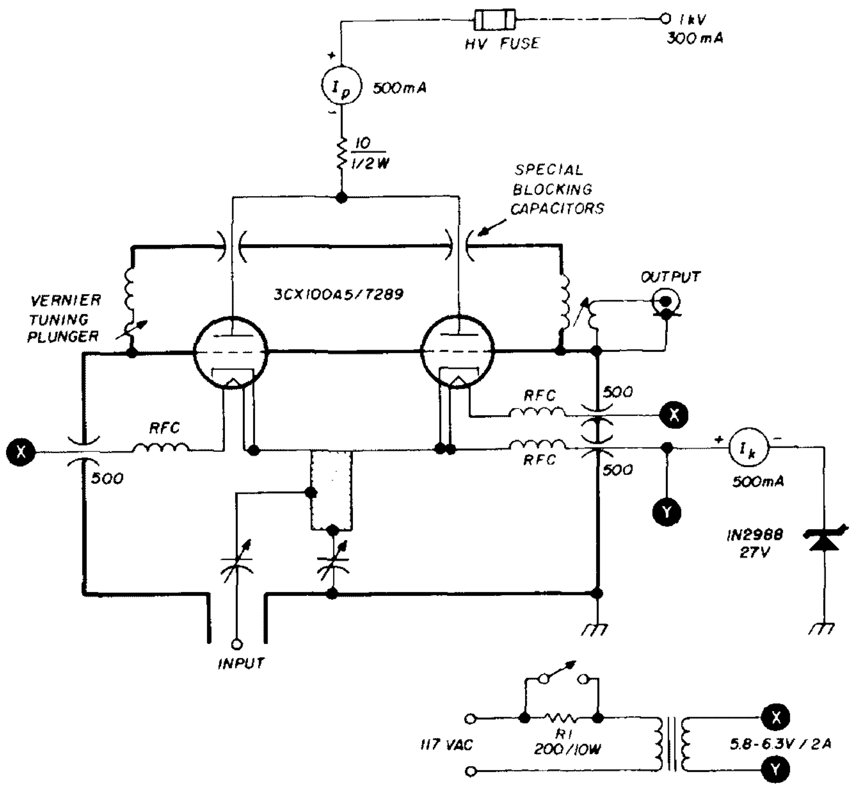



**fig. 1. Schematic of the dual 3CX100A5/7289 grounded-grid amplifier for 1296 MHz. Each RFC is 10 turns no. 22 solid copper wire on a 1/8-inch diameter form.**

pass capacitor, fig. 3, are made of 1/16-inch thick brass sheet. Finger stock, which contacts the tube plate rings, is soldered to the edges of the 1.18-inch-diameter holes in the top bypass capacitor plate. To facilitate alignment, the tube may be inserted into the structure during the soldering operation if a large soldering iron is used. Care must be taken to prevent solder from reaching the tube plate ring. The bypass capacitor plate is secured to the top cover with 8-32 Nylon screws. Teflon, 0.010 to 0.020 inch thick is used as the capacitor dielectric; although polyethylene, Mylar or mica will probably work as well.

lead screw, which is turned by a knob on the front panel.

The output coupling link is attached to a piece of homemade rigid coax (fig. 2), which comes through the center of the rear-cavity wall. The coax is retained by a shaft lock once the position for best loading is located. The rear wall should be temporarily clamped in place until cavity resonance is assured, then it may be permanently fastened to the top and bottom cavity plates with several small machine screws. Note that the ends of the tuning plunger and the rear wall *do not* contact the side walls at the corners of the cavity. Electrical contact is not nec-

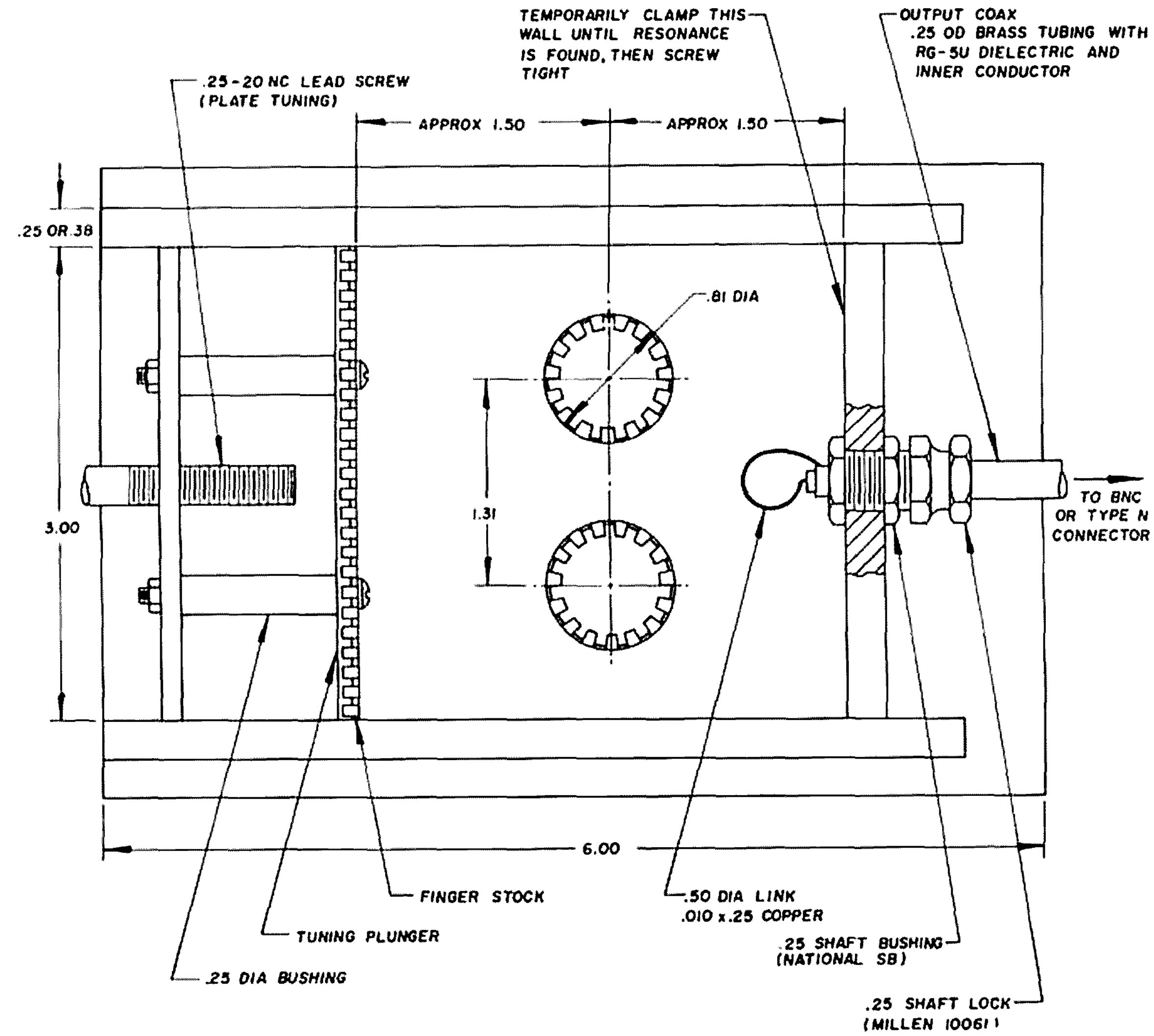
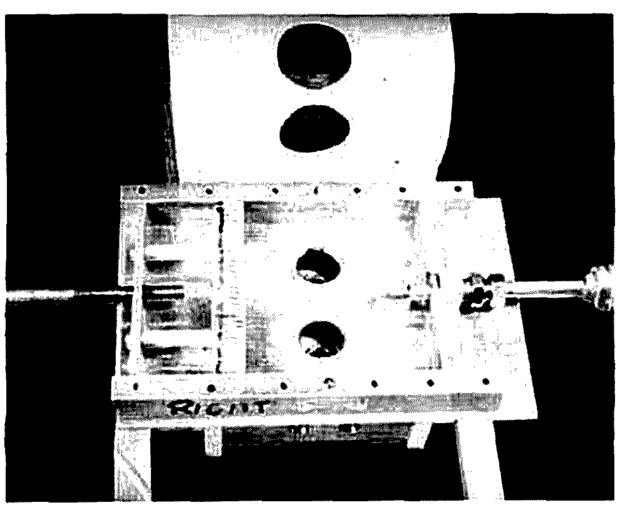



**fig. 2. Construction of the plate cavity.**

essary since essentially zero current is present in these regions.

Fig. 4 shows the cathode enclosure, which has walls constructed of ¼ x 1-inch brass rod stock. The stripline center conductor, which contacts both cathodes, is made from 1/16-inch copper and is supported by a single ½-inch-long ceramic standoff located midway between the two tubes. A 1/16 x ¼-inch strip soldered to the strip line forms one end of the tuning capacitor. The adjustable plate of the cathode tuning capacitor is the end of a threaded 3/8-inch panel bushing soldered to a ¼-inch-diameter brass rod, which is in turn coupled to a front-panel tuning knob via two universal joints. The small rf heater chokes are connected to the tube heater contact by small clips made of thin brass strip or pieces of finger stock material. The small spring strip is bent into a U shape and pressed into the heater contact.

Fig. 5 shows a cross section of both cavities. Input coupling to the cathode



**This view of the anode cavity with the top wall removed shows the output link and finger-stock pluger.**

cavity is provided by a probe that consists of a 5/16-inch-diameter brass disc soldered to the end of a threaded BNC connector. Not shown are several ¼-inch-diameter holes in the side walls of both cavities to facilitate air cooling of grid and cathode seals. A bakelite air duct covers both tube anodes. The blower air plenum should be arranged so that air will pass through the duct and also through the holes in the cavities. Note particularly the bushing at the tube anodes, which limits the insertion of the tube into the cavities. The tube should seat as shown by fig. 5 to preserve cavity resonance.

## tune-up accessories

Before applying power to the amplifier, you'll need a few simple accessories.

Substitute an adjustable resistor of about 200 ohms for the zener. This will be used to adjust cathode bias during tune up. You'll also need a 50-ohm dummy load and some kind of relative power indicator. If you don't have a 50-ohm load for 1296 MHz, a very good substitute can be made from sections of coaxial cable. Fifty feet of RG-8/U followed by 100 feet of RG-58/U will provide about 27 dB attenuation. This assembly will simulate a matched 50-ohm load at 1296 MHz.

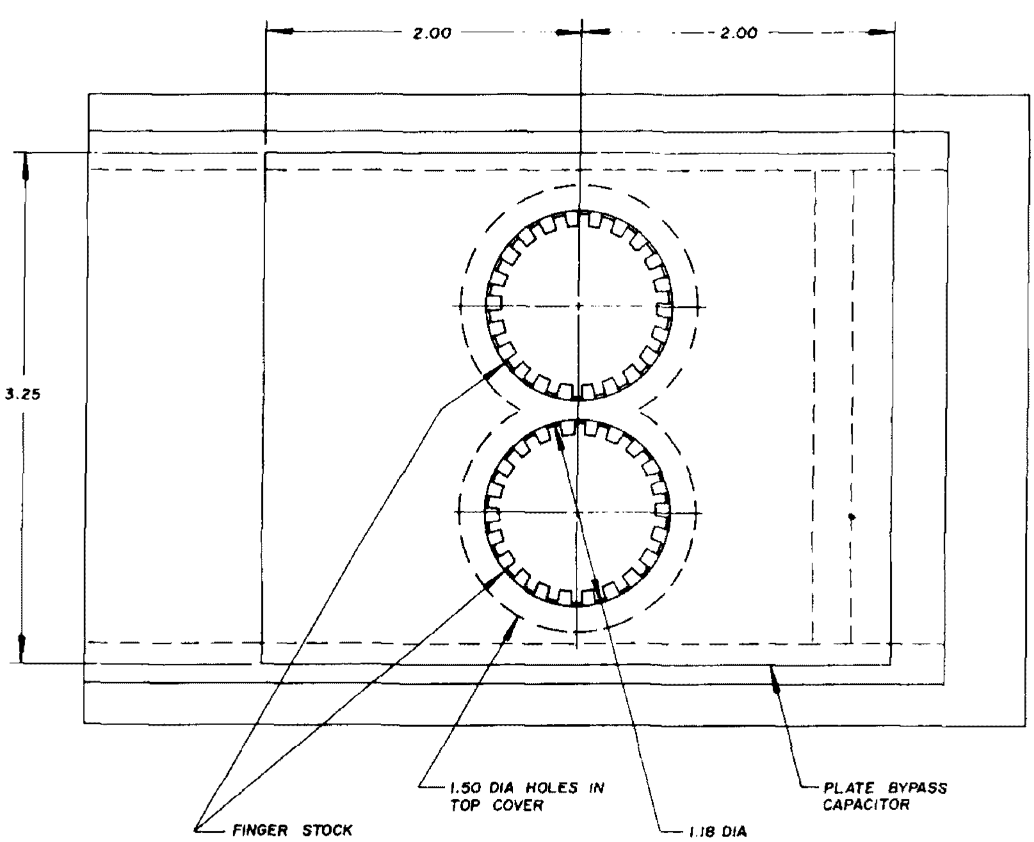
A diode detector and voltmeter circuit, such as that shown in reference 2, can be used as a power output indicator. The indicator circuit should be loosely coupled to the end of the dummy load. It's not necessary to calibrate the rf voltmeter since an indication of relative power is sufficient for initial adjustment of the amplifier. The coax sections may be coiled and taped.

Use only BNC and type N connectors. Don't use uhf connectors, as they will introduce a mismatch into the load.

## tuning up

With the blower running, dummy load and indicator connected, and no rf drive, apply about 500 Vdc to the amplifier anodes. Adjust the resistor for about 50 mA of anode current. Next, apply about 5 watts of drive power. Adjust cathode

fig. 3. Layout of the plate cavity top cover and bypass capacitor.



tuning for maximum anode current. Input and output coupling may now be adjusted for maximum output consistent with minimum dc input power. Bear in mind that coupling adjustments will alter the resonance of the tuned circuits. Therefore, both input and output circuits should be retuned only with optimum coupling.

Under these conditions, output power will be about 100 watts minimum.

## preventing self-oscillation

Unless a 50-ohm load remains on the output, the circuit may self-oscillate due to the high unloaded Q of the anode circuit. If an antenna switching device is to be used in which amplifier output is not terminated during receive periods, anode voltage should be removed before the antenna switch is operated. The anode voltage should be reconnected after the antenna has been switched to the amplifier in a sequential manner.

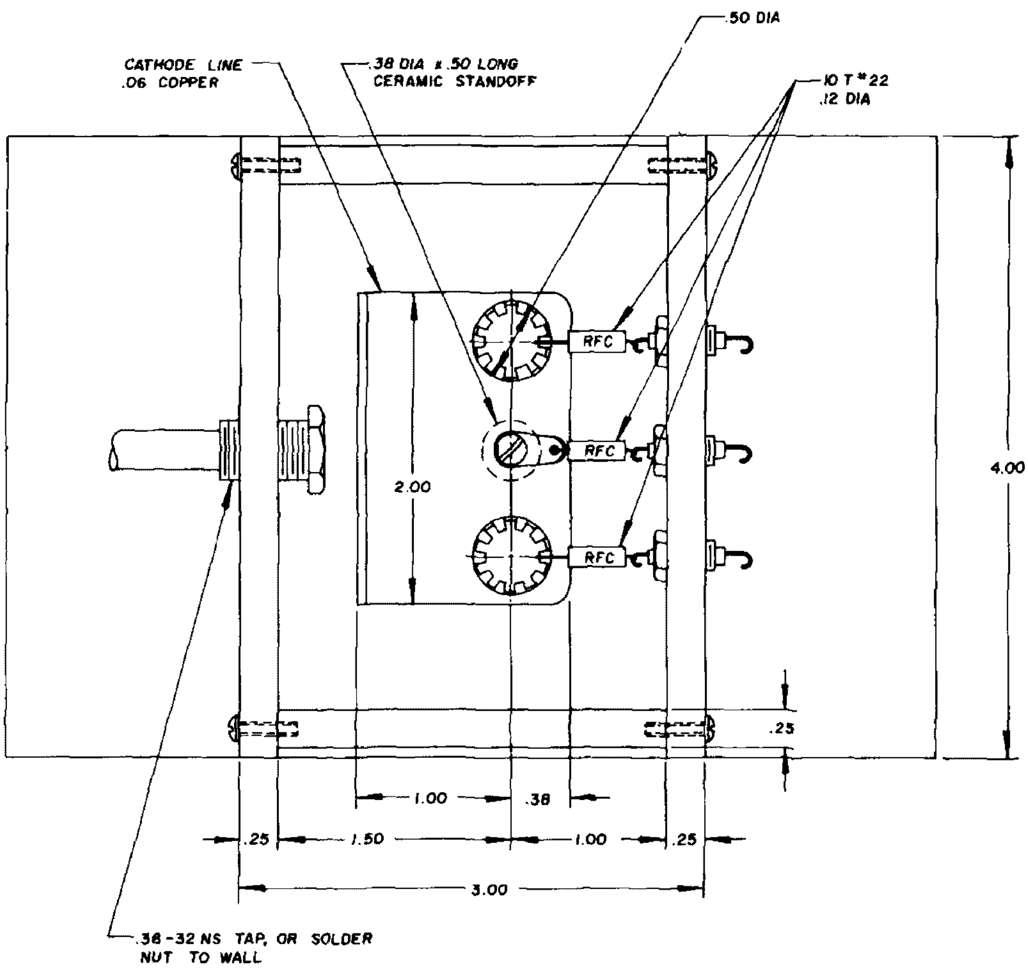



fig. 4. Cathode cavity of the 1296-MHz amplifier.

After the amplifier is optimized (40 to 50 percent anode efficiency), replace the zener. Apply full anode voltage (1 kV) and check tuning and coupling for optimum with about 10 watts of drive power. Grid current should be 50 mA if sufficient drive is available and all circuits are optimized. The tubes should not draw anode current when rf drive is removed.

With 1 kV on the anodes, the amplifier should load to 200 mA (anode) and 250 mA (cathode) with 10 watts of drive.

## tube dissipation

A quick check of the tube specifications indicates that a single 3CX100A5 is capable of 100 watts of anode dissipation, but our operating conditions call for

only half of this. We are being conservative for several reasons. The primary reason is to obtain reliable, long-term life in view of the high cost of the tubes (approximately $20 each). Amplifiers of this design have been operated with anode voltages up to 1.2 kV and output power up to 175 watts for short periods. Although rated anode dissipation is still not exceeded, grid and cathode heating can become severe should the amplifier be mistuned, lightly loaded, or over-driven.

We also found that, at the higher-power operating conditions, heater power could be removed entirely as long as the amplifier was operated continuously. These results point out an interesting property of most uhf planar triodes; maximum output is not usually limited by anode dissipation but more often by grid dissipation, since it is more difficult to extract heat from the relatively frail grid structure. A close-fitting grid ring connection with as much thermal conductivity as possible should be used. In addition, as we have done with this amplifier, vent holes should be provided next to the grid ground plane. This allows forced air to be applied directly to the grid and cathode seals, which is highly desirable.

## heater power

During operation we found that at 100 watts output, full heater power could be maintained with no serious consequences. However, it's recommended that heaters be operated at 5.8 instead of 6.3 volts. This will maintain adequate cathode temperature while minimizing back heating of the cathode during long standby periods. A more sophisticated method might be to include a relay that switches heater voltage under control of cathode current. Thus, when no drive is present, and cathode current is zero, the full heater voltage could be applied.

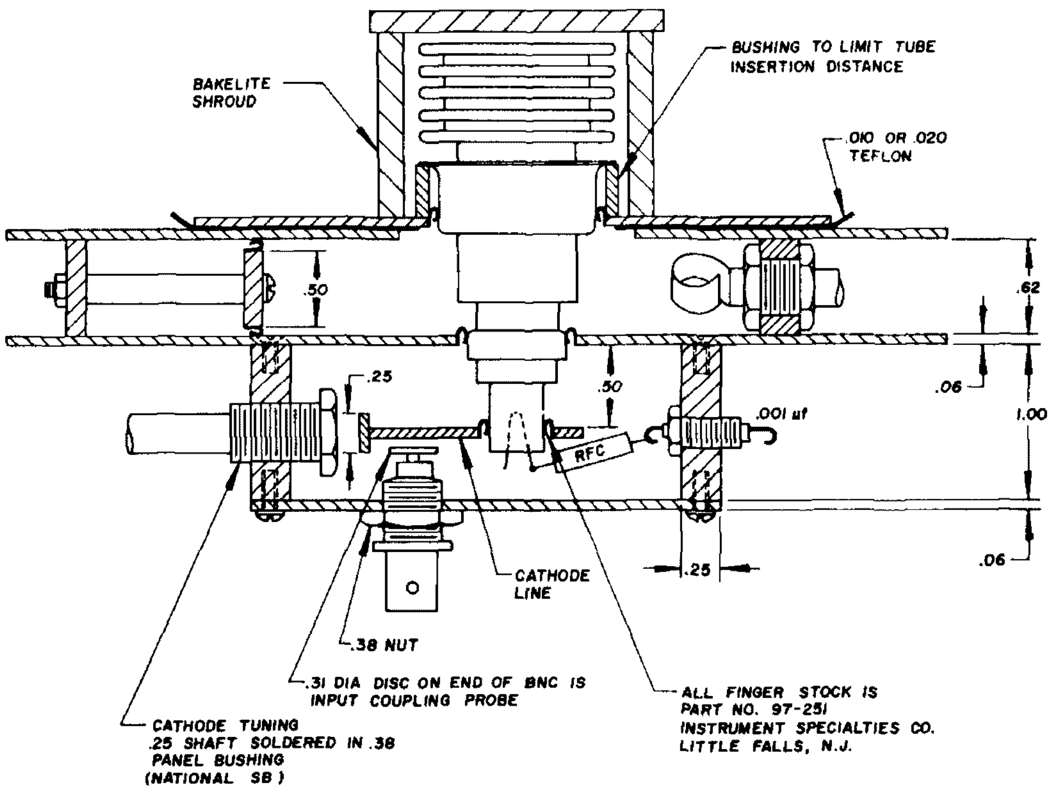



**fig. 5. Cross section of the amplifier, showing the plate cavity, cathode cavity, plate bypass capacitor and 3CX100A5 tube.**

## the dc blocking capacitor

The plate dc blocking capacitor deserves special mention, since its design is *not* based on capacitance alone. This capaci-

tor is considered as a flat parallel-plate transmission line. Its dimension in a radial direction from where dc blocking is desired is adjusted for a quarter wavelength or less. Thus, the physical length of the capacitor is reduced by the inverse square root of the capacitor's dielectric constant.

For example, Teflon has a dielectric constant, $\epsilon$ , of about 2.1. The radial length, L, of a dc blocking capacitor is then

$$L = \frac{\lambda}{4\sqrt{\epsilon}} = 9.14/4\ (2.1)$$

$$= 1.6 \text{ inches or less for 1296 MHz}$$

Additionally, the parallel-plate transmission line's characteristic impedance should be as low as possible. This is why a dielectric is used to load the line and to provide uniform close spacing. The reasoning is that, if the blocking capacitor looks like a quarter-wave section of transmission line, the end outside the rf cavity will be open. It will then reflect a very low impedance back into the cavity where blocking is desired.

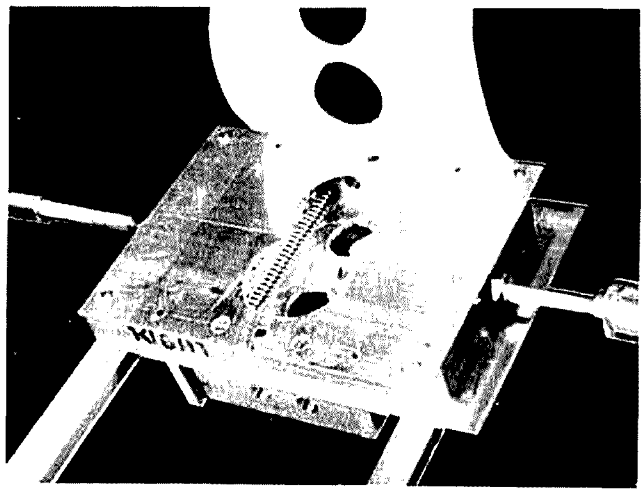
You might ask the question as to how much rf voltage appears at the "hot" end of the quarter-wave section outside the cavity. This is simply the product of the rf current entering the line and the line's characteristic impedance, $Z_0$.

Typically, $Z_0$ for a close-spaced parallel plate line with dielectric loading will be a few ohms. The current into the line is determined by where the line ends in the cavity. It is most desirable to place the dc blocking capacitor at a high field point where rf current is minimum. In most cases, this is impractical, and the block is placed where some current flows. However, even if the current is of the order of several amperes, the product of current and $Z_0$ would still be about 10 volts, a not too objectionable level. This approach to dc blocking in uhf amplifiers is very practical for most applications.

## concluding remarks

Note that very little soldering is required in assembling the cavity and enclosures, except for securing finger stock. Tight joints where necessary and mechanical stability are provided by the large area overlaps of the bar and sheet stock. This method of construction is less tedious but more costly.

The glass-sealed 2C39 tube, which is more commonly available, has been tried in this circuit with inferior results. Gain, efficiency, and output power suffer by at least 3 dB, and no attempt should be made to obtain more than 50 watts output with these tubes. A number of tubes failed because of punctures and cracks in the glass seal between grid and anode due to rf heating losses in the glass.



Anode cavity top wall with the anode plate removed. The teflon dielectric for the blocking capacitor is at the top.

For those who need a high-power tripler to operate from an existing 432-MHz transmitter, it is suggested that the same amplifier be modified in the cathode circuit to resonate at 432 MHz. This can be done in several ways. Perhaps the most simple way is to increase the stripline inductance until resonance at 432 MHz is achieved. Output power of about 40 watts should be readily obtained at reduced efficiency.


1. P. Laakmann, WB6IOM, "Cavity Amplifier for 1296 MHz," QST, January, 1968, p. 17.
2. The Radio Amateur's Handbook, ARRL Staff, 1969 edition, p. 547.


ham radio

# low-power solid-state transmitter for two meters

The will to improvise, plus salvaged tv parts, resulted in this little bomb

W. G. Eslick, KØVOY, 2607 East 13th Street, Wichita, Kansas 67214

I wanted a portable 2-meter transmitter, so I built one described in another magazine. I wasn't satisfied with its output nor with the critical tuning to obtain upward modulation, so I decided to "roll my own." The rig shown here is the result.

It produces 250 mW output measured on a Bird wattmeter. Unfortunately, I wasn't able to use inexpensive transistors in all stages. After trying several types in the final amplifier, the only device that provided upward modulation was the 2N3866.

## construction

The schematic is shown in fig. 1. I'd suggest staying away from PC boards on this band. I built several 2-meter units on PC boards, even on ground lines. A metal chassis is recommended for this rig.

I made the coils from old i-f transformers salvaged from a tv set. The transformer is the type whose slug has a hex slot (for a tuning tool) instead of a threaded brass screw. Most of the resistors and capacitors can be scrounged from tv sets as well.

The slug for the doubler coil was

sawed in half to obtain sharper tuning. If the unit won't load into your antenna, add a 10- to 20-pF capacitor across C10.

## cooling

Heat sinks are available most anywhere. Mine were scrounged from old IBM circuit boards. Wakefield fin types are fine. Mine is the slip-on type, which is held to the transistor with three set screws. I once blew the modulator transistors by using them without heat sinks. I replaced these with 2N1374's from some surplus boards from Radio Shack, and they work fine.

grid dip oscillator. The output from the 72-MHz stage and that from the doubler (145 MHz) was sufficient to peg the meter on my Millen gdo when in the diode position.

A trick described in reference 1 is interesting. A resistor connected between B+ and Q3's base will change Q3's operating angle. For example, a 10K-ohm resistor will make Q3 operate at something less than class C, or close to class AB. I didn't find this necessary for proper operation of the transmitter, however, I mention it for those who might wish to experiment.

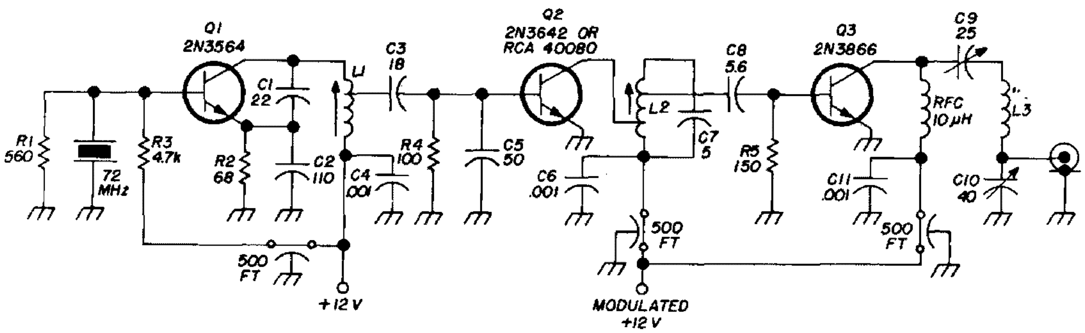



fig. 1. Schematic of the two-meter transmitter. A metal chassis is recommended, rather than a printed-circuit board, to avoid rf feedback problems. L1 and L2 are modified tv i-f transformer coils. A a-m modulator should provide at least 125 mW of power to adequately modulate this little rig.

L1 4 turns no. 20 bare tinned, spaced 1 wire diameter, tapped 3/4 turn from hot end

L2 4 turns no. 20 bare tinned, spaced twice wire diameter. Collector tap 1½ turns from cold end; C8 tap 3/4 turn from top end

L3 5 turns no. 18 bare tinned, ½-inch ID x 9/16-inch long

## the modulator

A suitable modulator for this transmitter is the Birnback modulator available from Round Hill Associates*. Other modulators that provide at least 125 mW of audio power can also be used. On a 6-meter transmitter, I had a hard time keeping rf out of the modulator. A clue to this was a low audio howl from the modulator output transformer, and the transistors ran very hot. Shielding, rf chokes, and bypass capacitors were required to tame the six-meter rig.

## adjustment

Each stage should be checked with a

## performance

With the transmitter in the basement, and a type 49 lamp as a dummy load, I worked as far as two miles fairly well. A whip antenna increased reports. Audio reports were crisp and clear.

One thing that might have caused the modulation problems with different final-amplifier transistors is that the modulator's 500-ohm output impedance is too high. But in homebrewing, you try to use what's available in your junk box.

With the transistors shown in the schematic, all worked well. One thing that should be remembered: connect the power backwards, and goodbye transistors.

ham radio

*Round Hill Associates, 434 Avenue of the Americas, New York, New York 10011.

# economical beam

## for ten meters

### Improving the "Wonderbar" antenna for effective DX work

R. A. Clymer, W1FPF, Pine Island Road, Mattapoisett, Massachusetts 07239

You've heard this old saw many times, but it bears repeating: without a beam antenna, it's futile to compete seriously for S/X. I was listening to everybody working all the goodies on ten meters recently and recalled the antenna I used in the mid-fifties during the last sunspot cycle peak. It was a simple piece of plumbing that *allowed you to get on ten* meters in a hurry. In its basic form, it's called a Wonderbar, or bow-tie antenna. Why not try it again, only this time crank some gain into it?

For those who may have forgotten, or who have recently joined the ham fraternity, I'll describe the procedure I used to adapt the original design[1] into an inexpensive beam for ten meters.

## construction

The main source of material was an old biconical tv antenna. My beam was modeled after the original Wonderbar design using these materials and some hardware from my junkbox. The basic Wonderbar antenna that resulted in shown in fig. 1.

I dismantled the tv antenna completely. I cut two 30-inch cross bars from the old elements. Each end of the cross bar was flattened and drilled to accept 3/16-inch machine screws (I used a 13/64-inch drill). Next, the crossbars were attached to the open ends of each of the two elements. This forms a couple of isosceles triangles, or wing-shaped elements.

I used a handy piece of 3/4-inch pine board, 13-inches long by 10-inches wide, for the base. Any material can be used that's sufficiently rigid to hold the assembly. Standoff insulators, female coax connectors, and a length of 5/8-inch OD heavy-wall plastic tubing (for spacers) were produced from my junkbox.

## assembly

Place the wing-shaped elements on the floor over the base. Space them about three inches apart. Drill six 13/64-inch holes (fig. 1) through the elements and completely through the base. (This will ensure alignment during final assembly.) Mount the female coax receptacle as

shown near the top of the base. Place the standoff insulators and spacers as shown, and assemble them loosely with 3/16-inch machine screws. Don't tighten the screws on the standoff insulator where the coil will be attached.

## the loading coil

If you don't have a B&W 3013 miniductor handy, it's easy to wind your own

1-3/4-inch form for the primary. This coil was removed from the form and slipped over the loading coil.

Attach the loading coil to the elements by securing each lead to the screws at the apex of the V formed by each element. Place a solder lug between the head of the screw and the element. Now solder the primary coil leads to the coax connector. Tighten all machine screws.

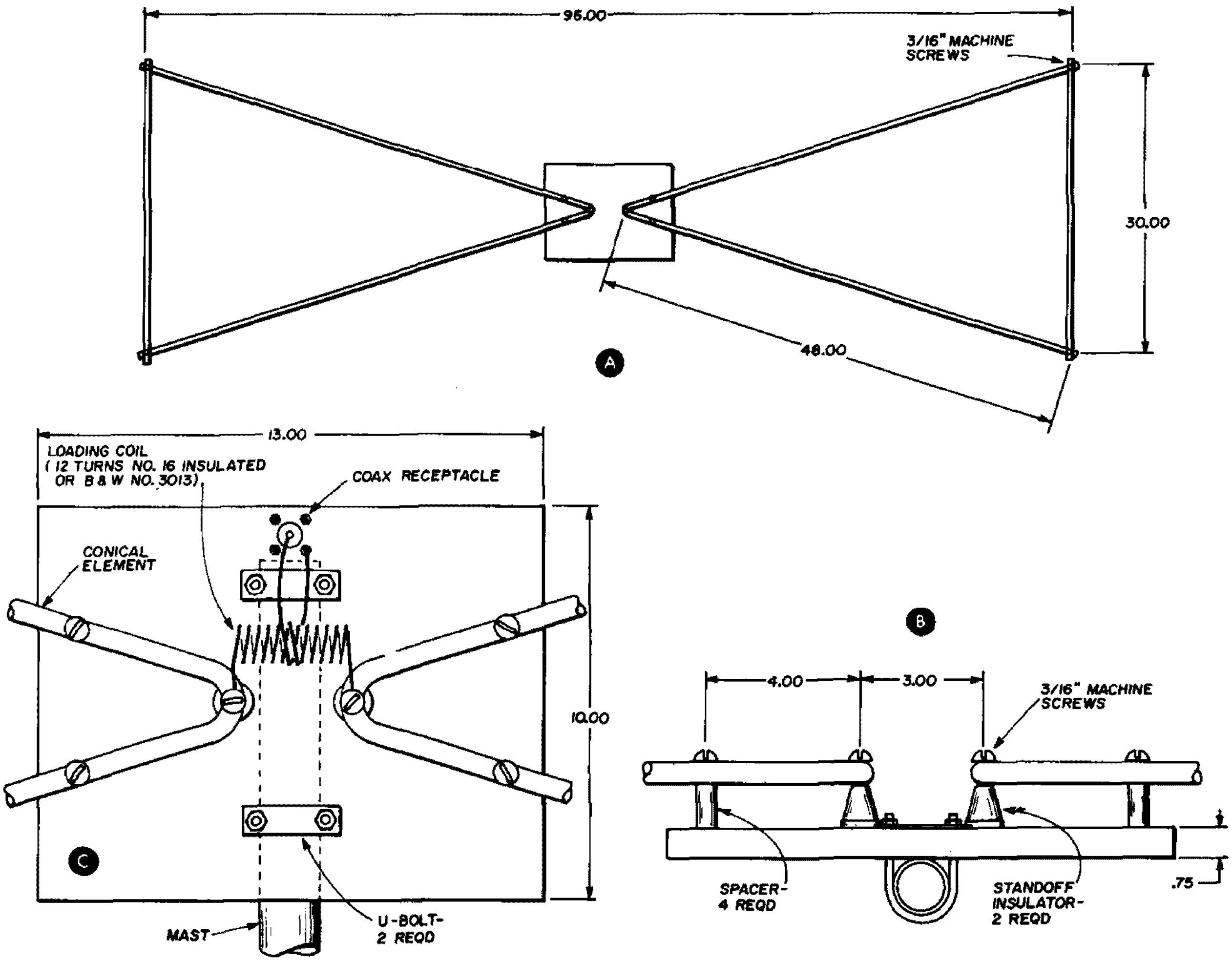



fig. 1. Dimensions, A, and mounting details, B, of the Wonderbar beam driven element. Mounting base is a simple pine board.

coil. I used what was available: number 16 insulated solid copper wire. Considerable latitude can be used here. Just make sure the coil is sufficiently rigid to be self supporting. I wound my loading coil around a 1-inch dowel, using 12 turns, close spaced. Then I removed the dowel and stretched the coil until it was about 3 inches long. Next I wound 2 turns of number 12 solid copper wire around a

## mast mount

Four holes are drilled to accept U bolts, which will secure the antenna to the mast. I used 5/16-inch holes, positioned over the centerline of the base. The first two were immediately below the coax connector, and the second two were about one-half inch from the bottom of the base. Use your own ideas here to fit your available hardware.

## preliminary tests

I attached a ten-foot piece of 1-1/4-inch conduit to my antenna for initial tune-up. I raised this assembly, with a piece of RG-8/U coax attached, in a vertical position and firmly lashed it to a picnic table. I found the best loading by tapping down on the loading coil; 10-1/2 turns seemed to be optimum. I made a permanent connection at this point by soldering. The lowest standing wave ratio (about 1.4:1) occurred at about 28.95 MHz.

With 65 watts input, I made contacts with two WØ stations, and got 5-9 plus reports. The next day, I worked a KV4 and a couple of G's. This simple antenna did indeed put out a good signal. But I wanted it to put out a better signal, so I added a reflector.

## the wonderbar beam

At this point you can enjoy this inexpensive antenna without further embellishments. It will provide a good signal on ten meters, it doesn't cost much, and you'll work some DX. However, if you like to experiment a little, as I do, you'll want to improve its performance. A simple reflector placed behind the Wonderbar antenna will produce from 3 to 5 dB gain over a reference dipole. This will effectively double your radiated power over the Wonderbar alone.

Handbook data showed that the shortest spacing for a reflector to improve performance was 0.15 wavelength (a little more than 4 feet on ten meters). This meant I could use the boom from the old tv antenna by merely adding a short extension. I decided to depart a bit from convention, for ease of assembly, and attached the boom and reflector immediately below the point where the Wonderbar was attached.

I made the boom extension about 5 inches longer than required. Then I put a bend of approximately 110 degrees radius in the extension about 4 inches from the end. I drilled two holes through the shorter leg of the boom extension and through the mast. This made an easy means of attachment. You could use the more conventional method of attaching the driven element and reflector at opposite ends of a one-piece boom. It would look prettier, perhaps, but wouldn't work any better. I wanted to use the materials on hand, so I used a short extension on the old tv boom.


"I'll go over in a couple of days and see how the beam held up..."

## the reflector

This element is simplicity itself. I cut my reflector from the remaining pieces of the old tv antenna tubing. It is 98 inches long (fig. 2). I used a pair of sheet metal cross braces I'd stripped previously from the tv antenna boom to attach the reflector.

## how far, wonderbar?

That's it. A simple, low-cost beam antenna made from materials on hand. The whole thing cost less than ten dollars. Results; I've worked fifteen different countries, many of them several times, including a ZS6. All this was done with a 60-watt a-m transmitter and my Wonderbar beam only twenty feet above ground.

### references

1. E.T. Bishop, K6OFM, "The Wonderbar Antenna," QST, November, 1956.

ham radio

# a stable small-signal source for 144 and 432 MHz

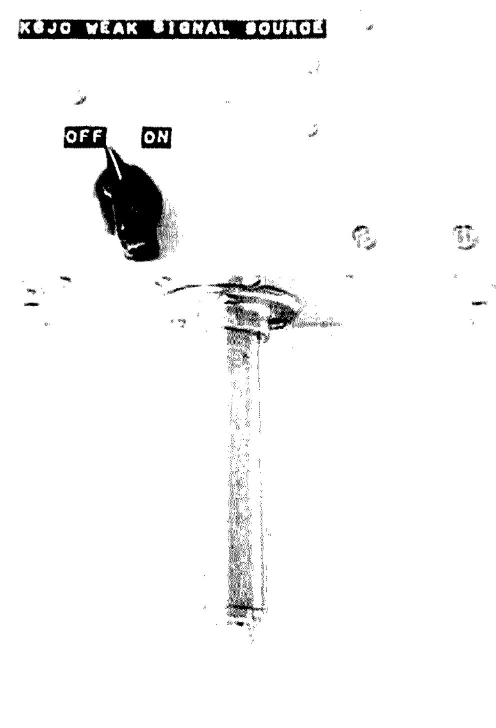
This simple circuit features variable frequency and amplitude control of a reference signal for vhf converter adjustments

James W. Brannin, K6JC, 424 Anson Avenue, Rohnert Park, California 94928

Have you ever made changes to your 144 or 432-MHz converter and tried to determine if the changes actually resulted in improved operation? Adjustments to converters on these bands can be disconcerting (and at times deceiving) without a stable signal source. The variable-output, crystal-controlled weak-signal generator described in this article allows immediate evaluation of converter adjustments as the work progresses. The signal is variable in amplitude on both hands, from several microvolts to below noise level. Frequency can be varied over a range of about 6 kHz on the 432-MHz band. The signal source can be calibrated against a commercial signal generator if one is available, although this isn't absolutely necessary.

## circuit description

The schematic is shown in fig. 1. A crystal oscillator, operating at 48 MHz in the emitter-base circuit of transistor Q1, triples to 144 MHz in the collector




Stable, small-signal source for receiver front-end adjustment. Close-spaced screws are necessary to avoid signal leakage.

circuit. A second stage, transistor Q2, triples to 432 MHz. The crystal holder should be grounded to obtain maximum tuning range of the crystal. The 52-ohm resistor in the attenuator should be non-inductive.

## construction

An LMB JB-880 or Bud CU-3006A Minibox may be used for a housing. These boxes have overlapping flanges that permit self-tapping screws to be placed about an inch apart. This seals the unit so no leakage exists.

The small subchassis is made from a 3½- x 2½-inch piece of aluminum with one edge bent for mounting to the top of the box with two 4-40 screws. The battery is held in place with a ½-inch aluminum strap fastened to the top of the box.

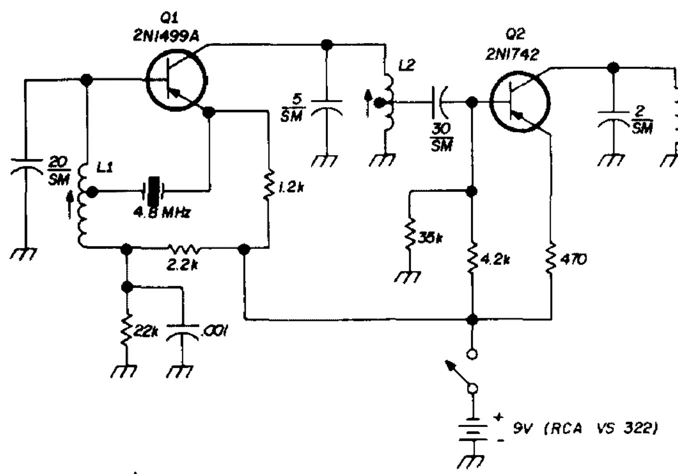



fig. 1. Schematic of the 144/432-MHz signal source. Frequency is adjustable over about 6 kHz on 432 MHz by slug in L1. Attenuator pickup look is spaced equal distances between L2 and L3. SM indicates silver-mica capacitor.

L1 11 turns no. 26 9/16" long, spacewound on 3/16" slug-tuned coil form, tapped 2 turns from bottom.

L2 7 turns no. 26 3/8" long, spacewound on 3/16" slug-tuned coil form, tapped 3 turns from bottom.

L3 2½ turns no. 26 5/16" long, spacewound on 3/16" slug-tuned coil form.

## the attenuator

The attenuator (fig. 2) is made from two pieces of brass tubing. The BNC connector may be sweated or screwed into the end of the smaller section. The cost to have the attenuator made by a machinist shouldn't be excessive.

Mount the attenuator on the side of the box so that the pickup loop is about the same distance from the 144- and 432-MHz tuned circuits. Output will be the same when using the signal source on either frequency.

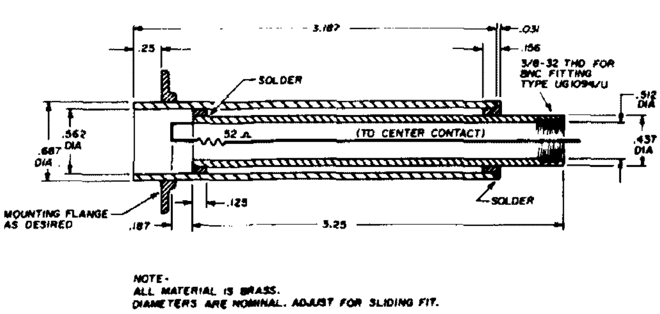



fig. 2. Construction details for the attenuator. Soft brass is used throughout.

## tuneup and adjustment

A vtvm with an rf probe should be used to peak the three tuned circuits. However, a grid-dip oscillator could be used. After the circuits are peaked, the frequency can be changed about 6 kHz on 432 MHz by adjusting the slug in L1. A receiver S-meter can also be used to peak the tuned circuits if the signal source is placed close enough to the converter or receiver front end.

After alignment is complete and the unit is carefully sealed with the self-tapping screws, practically no drift will be noted.

My thanks to W6PBC for the circuit design and to W6SPB for building the attenuator.

**ham radio**

# regenerative detectors
## and a
# wideband amplifier
## for experimenters

**Easy projects to acquaint you with transistor circuits, with hints on determining component values and some good advice on power-supply design**

Bill Wildenhein, W8YFB, RD 2, Blanche Avenue, Elyria, Ohio 44038

When testing converters, i-f strips, and similar equipment, it's desirable to have a simple "tail end" that covers a fairly wide frequency range, includes a means for receiving ssb and cw signals, and has good sensitivity. Another handy piece of equipment to have around your station is a general-purpose wideband amplifier. Presented here are some circuits I've found to be useful for experimental and general shop work. They're solid-state adaptations of proven, reliable circuits that have been around a long time. If you've never experimented with transistors, these projects will provide a good starting point to get acquainted with the fascinating world of solid-state devices. The circuits are simple, easy to build, and parts are quite inexpensive.

### regenerative detectors

For occasional bench work, I had been using a little regenerative receiver that dates back to the days of the big Kurz-Kasch dials. With only two tubes, it was completely adequate for the job. But one day I became annoyed at having to hook up both a vacuum-tube power supply and a solid-state supply. An era of profanity prevailed as I tried to get common transistors to perform as well as the old type 75 tube. Finally, the circuit of fig. 1 was devised. Greatest gain occurs with fairly heavy drain current on Q1 (R3 almost zero). The source tap should be adjusted for the point nearest the ground end of the coil that will sustain oscillation.

The circuit of fig. 1 worked quite well. The original haywire breadboard setup picked up shortwave broadcast stations from all over the world on 32 meters.

Next, I tried a version using the Colpitts feedback principle, fig. 2. I found that the antenna could be connected to the source instead of the gate. This resulted in lighter coupling to the antenna with fewer dead spots as compared to the Hartley circuit. I used a short length of RG-174 coax to feed the signal to the source. The coax capacity is in parallel with C8, which makes a convenient means of isolation. An advantage of this circuit is that feedback can be adjusted by C7 to obtain optimum performance.

## crystal-controlled detector

Going one step further, a crystal was substituted for L1, and now we have a crystal-controlled detector (fig. 3). In this circuit, R10 provides a dc return for the gate. Don't try to use an rf choke here, because the detector will take off simultaneously at the crystal frequency and that formed by the rf choke in parallel with the capacitances in the gate circuit. I found that C8 will vary considerably due to circuit layout and gain of the fet. The best way to proceed is to set the feedback capacitor (C7) to its midpoint, then connect various values of mica capacitors at C8 to find one that just barely produces oscillation. C7 can then be adjusted to compensate for variations in antenna loading and supply voltage.

Old timers familiar with the "blooper" regenerative detector will wonder where the grid leak and its capacitor went. They aren't necessary; the circuits will oscillate vigorously without them. In fact, if one of the circuits is to be used as a "tail end", it should be packaged in a small minibox to minimize stray radiation.

Somewhat more gain can be obtained from the circuit of fig. 4. The source bias circuit (C3, R1) should be retained. This contributes to smooth regeneration control. As with the old bloopers, the gate leak (R11) should have a high value for maximum sensitivity – 3 to 8 megohms. However, this may be at the expense of increased crankiness. In general, all three detectors seem to perform best with minimum capacity at C2. The antenna may also be fed at the tap on L1 in the circuit of fig. 4.

## oscillators

The detector circuit may also be used as a general purpose oscillator. If you wish to use it as a crystal oscillator as in fig. 3, C8's value will largely determine frequency range. Appropriate values for C8 in this case are:

0.1 – 0.4 MHz, C8 = 0.005 μF
0.4 – 4.0 MHz, C8 = 0.001 μF
4.0 – 10 MHz, C8 = 100 pF

Oscillation occured at all three ranges with a 7-100 pF compression trimmer for C7. I found, however, that C7's value had to be increased to about 500 pF to obtain zero beat with WWV using certain 100-kHz crystals. The waveform was not sinusoidal and had high harmonic content; output was 6 volts peak.

## variable-frequency oscillator

The circuits of fig. 1 or fig. 2 may be used as a general-purpose vfo. However, I'd recommend using a Vackar in preference to these circuits. While acceptably stable, neither circuit will compare to a Vackar oscillator. I've used a vfo version

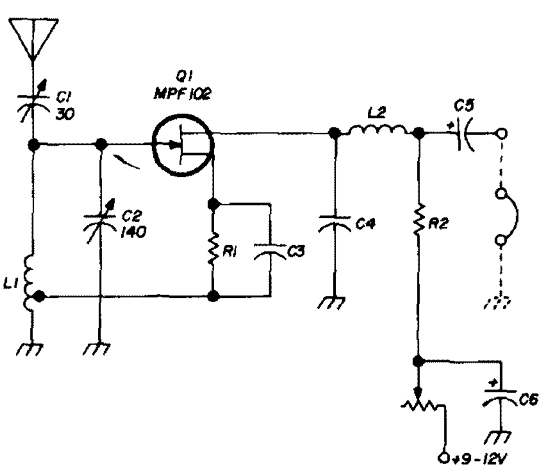



fig. 1. Original Hartley circuit used to replace a two-tube "blooper" for bench work. With a Motorola MPF 102 jfet, the set received shortwave broadcast stations from all over the world on 31 meters.

of the detector circuit in an experimental receiver. While it was stable as a detector in a 1700 kHz i-f amplifier, its amplitude stability over a wide frequency range was far inferior to the Vackar. Both this circuit and the Vackar have the same general ratio of maximum-to-minimum frequency range, but I can't recommend this oscillator as being competitive with the Vackar.

## audio amplifier

Detectors such as those shown can be improved with the addition of a high-gain audio amplifier. The amplifier shown in **fig. 5** is very useful around this station. It has quite good frequency response. **Fig. 6** shows what can be obtained with inexpensive epoxy-encapsulated silicon transistors. The amplifier will deliver a clean sine wave of 8 V p-p at frequencies from less than 10 Hz to at least 8 MHz.

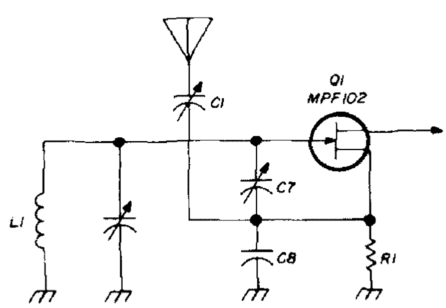



**fig. 2. Regenerative detector using the Colpitts circuit. Antenna can be connected to transistor's source, which results in fewer dead spots across the tuning range.**

This should be sufficient warning that decoupling the supply voltage may, in some cases, be necessary to prevent "motorboating" or other evidence of undesirable feedback. If you use a stiff, well-regulated power supply, or batteries, this shouldn't be much of a problem. However, if the power-supply impedance is too high, you will have problems (more of this later).

## amplifier applications

I've built several of these amplifiers and have found many uses for them. For example, many excellent solid-state vfo's are limited to less than a volt output. This amplifier will boost the vfo output to several volts without additional tuned circuits. It will also make a good i-f amplifier.

If you experiment with filters, your vtvm possibly won't have a range low enough to allow measurements below about 40 dB. One of these amplifiers, well shielded and battery operated, can extend the average vtvm's range to about 1 mV full-scale. However, it's well to calibrate the amplifier to be certain of its gain linearity. The response isn't flat, but this won't matter if measurements cover a limited frequency range, such as that of an ssb filter, for example.

Still another factor to consider is the amplifier's impedance. It won't compare to that of the usual vtvm. With a high beta transistor at Q2 in the amplifier, you can expect an input impedance of about 10 kilohms. Therefore, any high-impedance circuit (1 megohm or more) will be loaded by this amplifier. However, it's adequate for most solid–state measurements.

If the amplifier is overdriven, it will clip, producing a pretty good square wave. If it's used as a vtvm amplifier, it should be checked on an oscilloscope. This will indicate the maximum voltage available to your vtvm before the amplifier's waveform becomes degraded. Stay well within the maximum voltage range. Amplifiers I've built show that this voltage remains pretty constant throughout the amplifier's range. If you have an inexpensive oscilloscope, you can find

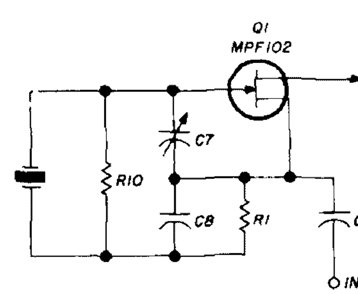



**fig. 3. The Colpitts circuit with a crystal substituted for L1. Capacitor C8's value will vary, depending on circuit layout and transistor gain; use a value that just produces oscillation.**

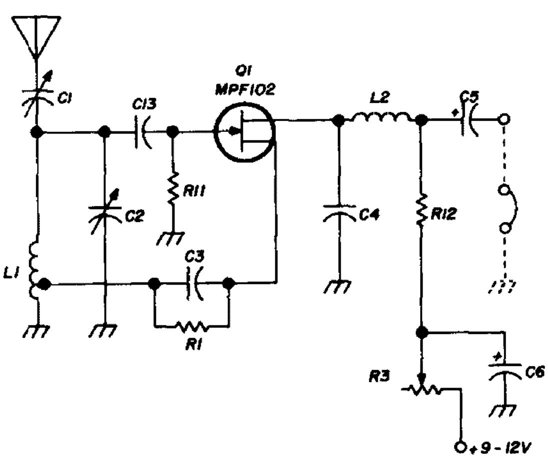
table 1. Parts list for the detectors and wideband amplifier.

| | |
|---|---|
| C1 | 3-30 pF trimmer |
| C2 | 140 pF variable |
| C3 | 0.005-0.02 µF disc ceramic or mica |
| C4 | 0.001-0.005 µF disc ceramic or mica |
| C5, C6 | 1-4 µF 25 V electrolytic |
| C7 | 3-30 pF compression trimmer or APC variable |
| C8 | 200-1500 pF mica (depends on fet, and degree of coupling) |
| C9* | 33 µF tantalum |
| C10* | 47 µF tantalum |
| C11 | 0.005-0.05 µF disc ceramic or paper tubular (see text) |
| C12 | 50-200 pF mica |
| C13 | 50 pF mica |
| R1 | 270-470 ohms ½ W |
| R2 | 1-1.5 kilohm ½ W |
| R3 | 2.5 kilohm |
| R4* | 10 kilohm ½ W |
| R5* | 1 kilohm ½ W |
| R6* | 100 ohms ½ W |
| R7* | 100 kilohm ½ W |
| R8* | 350 ohm ½ W |
| R9 | 10-25 kilohms |
| R10 | 47 kilohms-1 megohm ½ W |
| R11 | 2.2-6.8 megohms ½ W |
| R12 | 1-4.7 kilohms ½ W |
| L1 | 5/8 inch diameter, 32 pitch Miniductor 7/8 inch long, tapped 3 turns from ground end for 5-12 MHz range. (A 45 µH slug-tuned coil worked well in a 1700 kHz i-f strip.) |
| L2 | 2.5 mH rf choke |
| Q1 | MPF102 jfet |
| Q2, Q3 | 100 MHz plastic encapsulated silicon transistors. Dc betas: Q2, 200; Q3, 100. |

this point at, say, 60 Hz; then if you don't go above 50 percent of the point where distortion occurs, you can pretty well depend on the amplifier's response over its range.

## amplifier construction

Let's set up this little amplifier using




parts from your junk box. A complete list of parts for all the circuits described here is given in **table 1**. Because the amplifier is somewhat more critical as to transistor characteristics and resistor and capacitor values, I've included a discussion on how to grade your surplus devices and how to select resistors and capacitors to obtain optimum performance. The values in **table 1** marked with an asterisk were those I used in the amplifier whose response is depicted in **fig. 6**.

## grading transistors

Select only silicon transistors. With

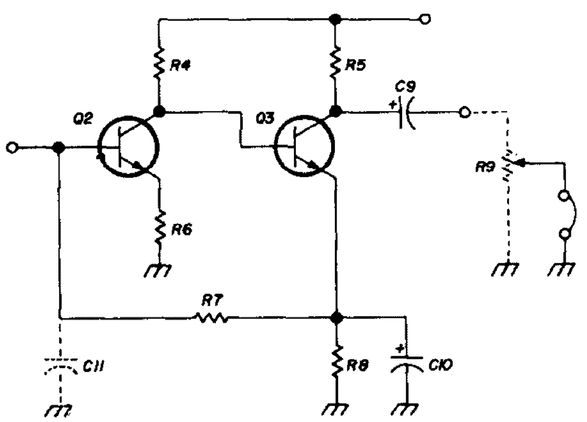



fig. 5. The detector circuits can be improved with with this wideband audio amplifier. Several applications of this basic circuit are discussed in the text.

high-gain, high-frequency types available on the surplus market at prices around four for a dollar, it's pointless to use less expensive germanium units with their higher leakage. As a rule of thumb, select transistors with a gain-bandwidth product, $f_T$ at least ten times as high as the expected pass frequency.

Illustrated in **fig. 7** is a circuit for using your vtvm to measure transistor current gain. Although this reveals little of a transistor's high-frequency gain, it will give a good relative indication of which surplus device is hottest. A simple method of grading your surplus transistors is as follows.

*Components used in an amplifier whose response is shown in fig. 6. See text for determination of values.

1. Set the 1-megohm pot to maximum resistance.
2. Set the switch to Ic.
3. Close the battery switch.
4. Set the 1-megohm pot to give a reading of 5 V dc on the vtvm.
5. Set the switch to Ib.
6. Note the reading in dc volts. (Assume it is 0.5V.)

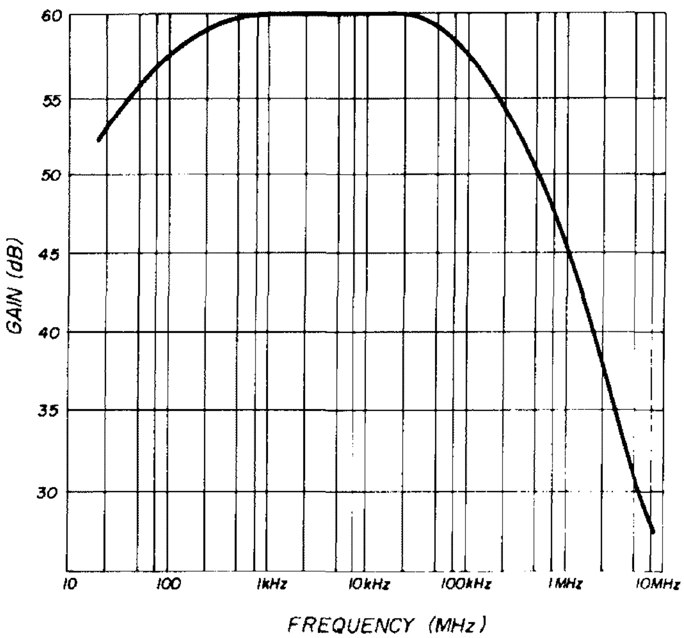



**fig. 6. Voltage gain as a function of frequency for the wideband amplifier. A pure sine wave of 8V p-p is produced over the entire range using inexpensive silicon transistors.**

7. Calculate the cd current gain:

$$I_c = \frac{E}{R} = \frac{5.0}{1000} = 0.005 \text{ A}$$

$$I_b = \frac{E}{R} = \frac{0.5}{10{,}000} = 0.00005 \text{ A}$$

$$G = \frac{I_c}{I_b} = 100$$

Using this method, you can grade transistors with dc betas between 5-500 with ease, which will give you a good idea of which transistor to use for Q2 in the amplifier. It reveals little about the transistor's ac characteristics, but it's a good starting point. Now select another transistor, which can have a lower beta, for Q3. We'll now "ballpark" the other component values.

## resistor and capacitor values

R5 is chosen to drop the collector voltage to about 50 percent with a chosen safe collector current. If the collector current is too low, gain will decrease. If it's too high, collector dissipation will be exceeded. A safe value for most surplus transistors is 5 mA. Using a 12-V power supply, we wish to drop about 6 V across R5:

$$R = \frac{E}{I} = \frac{6.0}{0.005} = 1200 \text{ ohms}$$

Next, R8 is chosen to drop about 10 percent of the power-supply voltage:

$$R = \frac{E}{I} = \frac{1.2}{0.005} = 250 \text{ ohms}$$

(270 ohms is adequate.)

Now the base voltage of Q3 is determined. Since Q3 is a silicon npn transistor, its base voltage will be about 0.5 to 0.7 V more positive than its emitter. From the above, we see that Q3's emitter is biased by R8 to about 1.2 V positive. Add the base-emitter voltage of Q3, and we find that Q3's base should be about +1.8 V.

Transistor Q2 can operate at a lower current to save total power. Let's say we'll operate it at 1 mA. The lower end of R4 is connected to Q3's base. We know that Q3's base will be 1.8 V. The total voltage drop across R4 is then 12 V minus 1.8 V. Current is 1 mA plus Q3's small base current, which can be neglected. R4's value then becomes

$$R = \frac{E}{I} = \frac{10.2 \text{ V}}{0.001 \text{ A}} = 10.2 \text{ kilohm}$$

A 10 kilohm resistor will be fine.

R6 provides a small amount of degenerative feedback. As its value is increased, the amplifier's input impedance increases and frequency response improves, but at a sacrifice in gain. Values between 100 and 330 ohms will be adequate for general use.

C10 can have a capacitive reactance of about one-tenth the value of R8 at the lowest frequency of interest. R8 was determined to be 270 ohms. If response to 100 Hz is desired, a reactance chart

(see the ARRL handbook) will give a value that represents about 27 ohms at 100 Hz; this will be about 60 μF. A 500 μF capacitor would be required to extend coverage to 10 Hz.

Large values of electrolytics cease to be bypass capacitors at these frequencies, so to obtain good high-frequency response it's advisable to parallel them with mica or disc ceramic capacitors. C9's reactance can be ten times as high as that of C10 in this circuit, so values of 5 μF or higher will be adequate.

## selection by substitution

R7 is best determined by experiment. After the entire circuit is wired except for R7, connect a 1 megohm potentiometer in place of R7. Connect your vtvm from Q3's collector to ground. Turn on the power, and adjust the pot so the voltage from Q3's collector to ground is about 6 V. Turn off the power, and measure the pot resistance. Choose a resistor that measures the same, and substitute it for the pot. The resistor value doesn't have to be precisely correct. If this fixed resistor produces a collector voltage of 5 – 7 V, the amplifier will operate all right, but permissible signal-voltage swing before clipping will be reduced somewhat.

When the amplifier is used following a regenerative detector, a volume control is desirable. A pot of 10 – 25 kilohms connected as shown (R9 in fig. 5) will be adequate.

## the importance of C11

C11 is added to the amplifier to limit frequency response. When used with a regenerative detector, high-frequency hiss is annoying. A capacitor of 0.05 – 0.1 μF at C11 will do much to make listening more enjoyable. It will also make the amplifier more stable in this instance. Even though the regenerative detector uses an rf choke and capacitor to filter rf components, rf can still sneak through to cause overloading in the amplifier. C11 will eliminate this.

Still another benefit of C11 is this: without C11, the turn-on surge of voltage from the detector drain is sufficient to cause momentary pumping in the audio amplifier. Voltages soar up and down drastically. In one such amplifier, high voltage was severe enough to zap Q2. With C11 in the circuit, these wild excursions are minimized.

## as a speech amplifier

This little amplifier will also make an excellent speech amplifier for a high-impedance microphone. Connect the hot lead from the microphone to Q2's base

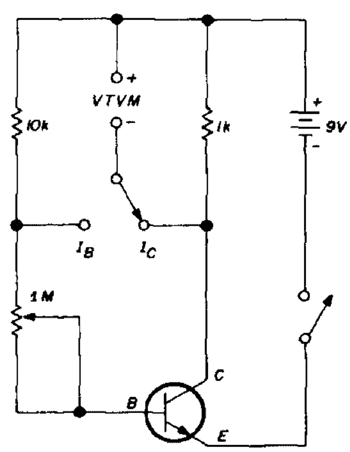



fig. 7. Circuit for using your vtvm to grade surplus transistors. It will give a good idea of which transistor is hottest.

through a resistor of about 27 – 47 kilohms. C11 is proportioned to serve as an rf filter; use about 0.01 – 0.005 μF.

## i-f amplifier application

In i-f amplifier use, R5 can be replaced with a tuned circuit. If so, it would be well to decouple Q2. Add a 270 ohm resistor from the top of R5 to the top of R4. Connect an 0.01 uF capacitor from the junction of the 270 ohm resistor and R4 to ground. C9 and C10 can be much smaller than when used in an audio amplifier. At 455 kHz, C10 can be 0.01 μF, and C9 can be 0.001 μF. These values will allow operation well into the megahertz region.

## power-supply impedance

This piece is written to encourage the vacuum-tube gang to experiment with solid-state circuits. An item I mentioned previously – power-supply im-

pedance — is worth some discussion.

With tube circuits, you may have seen many examples of what happens with poorly regulated power supplies. When you were servicing an ac-dc set, for instance, perhaps it broke into oscillation. Merely replacing the electrolytic in the power supply possibly solved the problem.

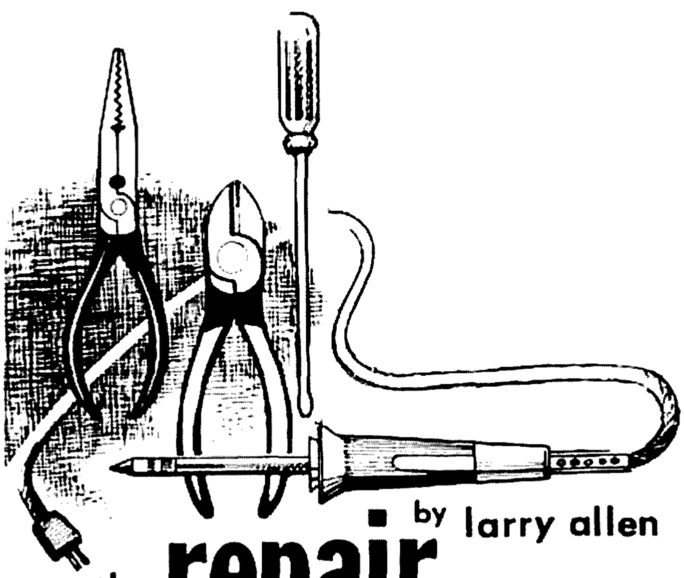
As another example, you may have had a modulator with push-pull 807's. If this modulator were fed from a "limp" power supply that allowed screen voltage to drop as you hit the mike harder, you may have found that by regulating modulator screen voltage you apparently picked up a lot of modulator power. Many high-gain circuits have been cussed endlessly, when in reality the poorly regulated power supply was the real culprit.

Poor regulation is even more critical in solid-state work. One simple solution is to use batteries. Fresh batteries do a very good job. Even better is a 12 V car battery. But the best solution is to build a properly regulated power supply for transistor work.

Many people ask "What do you mean — power supply impedance?" Suppose you have a power supply that's supposed to deliver 100 V at 200 mA. Maybe it delivers 100 V with no load, but at 0.2 A load, it skids to 75 V output. This means you've lost 25 V at 0.2 A. It's the same as saying the power supply looks like a good 100-V job with a 125-ohm resistor in series with the supply lead. What you want is, ideally, a power supply that will deliver the same voltage regardless of load — a "zero-ohm" power supply. You won't achieve this, but a properly built supply's impedance will be way down in the milliohm region.

I can't stress too highly that if you're just starting out in solid-state work, a well-built, regulated power supply will be one of your best investments. With it you can build high-gain circuits with far less bugs than would be possible with the limp and soggy supplies being used by many amateurs.

**ham radio**



# the repair bench

by larry allen

From what I hear, many amateurs consider an oscillator the hardest kind of stage to repair. That's too bad, because ham rigs are full of them. Maybe oscillators seem tough because of their bootstrap nature—making a signal out of what seems like nothing.

"If an oscillator quits," ask a good many hams, "just where do you begin looking for the fault?"

Instead of giving a quick, simple answer, I'll first explain what an oscillator really is. That'll prepare you for the easy testing methods I use. The truth is, an oscillator shouldn't be hard to fix.

## the four needs

Every oscillator has four requirements. Take away any one of them, and it just won't work. They're the key to an oscillator's operation—and also to its quitting.

1. An oscillator needs amplification. That's why the stage has a tube, bipolar transistor or field-effect transistor. The reason for this need becomes clear as you understand the second requirement.

2. An oscillator needs feedback. A small amount of signal from the output of the amplifier is fed back to the input. There, the tube or transistor amplifies it again. Then a sample of that output is again fed back. It's a sort of round-robin. Without this positive or regenerative feedback, the stage is only an amplifier.

3. An oscillator needs dc power. The tube or transistor needs dc voltages. Further, no amplifier is 100 percent efficient, and some power is wasted keeping the signal going round-and-round. The plate or screen-supply line in tube oscillators, or the collector or drain supply in transistor oscillators, makes up for this wasted power.

4. An oscillator needs frequency control. An oscillator is a signal-generating device. By its very nature, a signal has some frequency. Inductor-capacitor circuits or crystals set frequency in rf oscillators, and RC time-constant circuits do it for audio oscillators.

So there you have the four needs of any oscillator, no matter what its purpose. It is the method of developing these four requirements that distinguishes the various kinds of oscillators. Be assured, every oscillator has all four.

## some oscillators

Just so you know what I'm talking about, here are a few oscillator stages. The four requirements are met in each one.

The first is in fig. 1. One of the most simple and popular oscillators in ham gear is this crystal-controlled Colpitts.

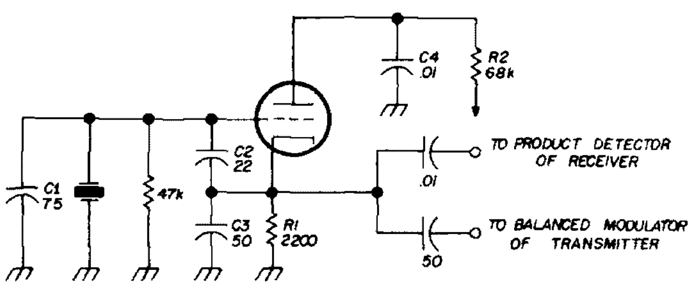
Amplification is by triode tube. It receives dc power through a resistor in the plate circuit. The cathode has a resistor that develops bias for the tube.

Feedback is by a capacitive divider connected between grid and ground. Notice that the plate is grounded for rf through C4. R1 is the load for the tube. Signal developed across R1 is divided up. Part goes to ground through C3, and a small amount goes to the grid through C2. The signal fed to the grid is amplified by the tube and developed across the cathode

load, and the process starts over. This regenerative feedback produces oscillation in the stage.

Frequency of the generated signal is controlled by the crystal from grid to ground. A crystal is common in communications rf oscillators, because of its extreme accuracy. The stage in **fig. 1** is called untuned because there is no LC circuit, but in reality it is tuned by the crystal. Without the crystal, there would be a coil in its place, resonated by C1 and the series combination of C2-C3. C1 or the coil would probably be made variable for tuning.

**fig. 1. Colpitts oscillator is used often in amateur radio equipment; it is distinguished by capacitive-divider feedback.**




The Colpitts configuration could be used just as easily with a transistor. In that case, the collector would probably be the element receiving dc power; the base would be connected where the grid is; and the emitter would replace the cathode.

A fet could be used, too. Its drain would take the place of the plate, its gate the place of the grid, and its source the place of the cathode. A Colpitts stage using a fet can be exceptionally stable.

One of the most popular crystal oscillators is the Pierce, and versions of it. Two of them are shown in **fig. 2.** The identifying characteristic of the Pierce is the plate-to-grid connection of the crystal. In that position, it supplies *two* of the oscillator's needs: feedback and tuning.

Dc power for the tube amplifier in **fig. 2A** comes through the plate load resistor. The variable capacitor lets you warp the crystal slightly, to pinpoint its frequency precisely. The capacitive divider in the output cuts down the amount of signal injected into the next stage; the vacuum-tube Pierce is a strong oscillator.

The version in **fig. 2B** is an electron-coupled Pierce. The screen grid (grid 2) of the tube is considered the oscillator plate. The crystal is thus connected from plate to grid in regular Pierce style. However, the generated signal is injected into the electron stream that goes to the plate, from which output is taken. The output load thus has little effect on oscillator frequency.

From what's already been said, you can imagine a transistor Pierce circuit. With a fet, it would look like **fig. 3A.** The resemblance between it, the bipolar transistor version in **fig. 3B,** and the triode in **fig. 2A** makes it look as if you could just plug in a transistor in place ot the tube. With a fet, you can; just reduce the plate voltage to a drain voltage that is safe. For

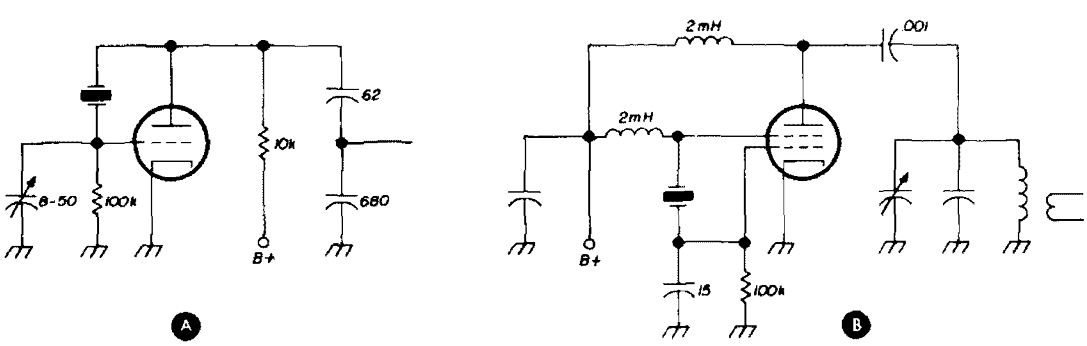



**fig. 2. Pierce oscillators are characterized by a crystal between the plate and grid.**

a bipolar, you must change resistor values, devise a base-bias system, as well as lower the supply voltage.

Keep in mind that all the oscillators I've shown you have those four requisites; amplification, dc power, feedback, and frequency control.

## the dead oscillator

What would lead you to suspect an oscillator of being dead in the first place? In a straight cw or a-m transmitter, that's pretty elementary. There just isn't any carrier when the oscillator quits. In an ssb transmitter, or in a receiver, the symptom might not be that simple. It all depends on which oscillator quits.

In one ssb transmitter, there are four different oscillators. One is the carrier generator. Without it, no sidebands can be produced, so the transmitter has no output. Unfortunately, other problems can cause the same symptom, so lack of output in a sideband transmitter doesn't prove the carrier oscillator is dead.

The other three transmitter oscillators are for heterodyning the sideband up to the transmitting frequency. Any one of them can block the sideband before it reaches the transmitter output stage. Usually, the tuned circuits that follow the frequency converter or mixer stage eliminate any sidebands but those of the desired frequency. Without correct oscillator injection, the mixer output can be nothing more than the sideband input from the preceding stages. Output is thus blocked.

A receiver using triple conversion can have as many as five oscillators. The unit diagramed in fig. 4 is an example. The calibration oscillator and the second version oscillator (feeding what's labeled the "first converter," but really is the second) are both crystal controlled. The third conversion oscillator is fixed in frequency; a stable Hartley stage is used.

The receiver is tuned to the incoming signal by the first conversion oscillator (feeding the "mixer"). It's a high-frequency variable oscillator that beats against the desired incoming sideband signal to form a sideband i-f at 3.035 MHz.

And, finally, there's the bfo. It's usually frequency-fixed in an ssb receiver. In a set that receives icw code transmissions, the bfo is generally variable over a small range so you can select the pitch of code signals you hear.

If any of the three conversion oscillators goes dead, you simply can't get a signal through the receiver. If the calibration oscillator stops, the only sign of it is a lack of birdies when you're checking calibration.

The bfo is what furnishes the injection carrier that lets the product detector recover the voice signals from the sideband. If it quits, you're likely to get a very weak and distorted sound from the phones or speaker.

## which one's dead?

The nature of an oscillator makes it easy to test, if you have the right equipment. I have a fet voltmeter, and I've

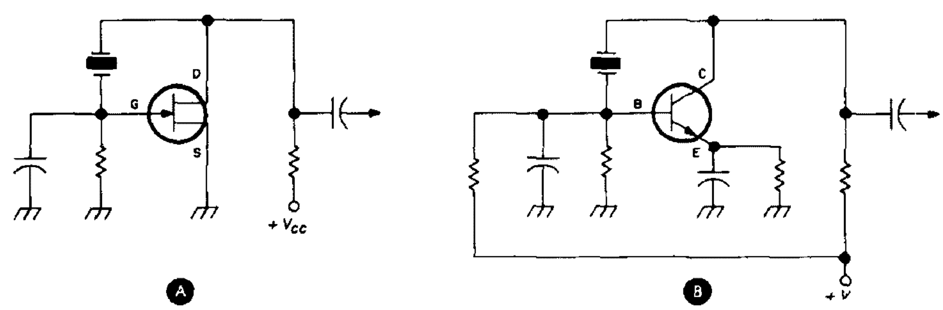



fig. 3. Solid-state Pierce oscillators using a field-effect transistor, A, and npn bipolar, B.

built a little probe that lets me measure rf. With it, I can find out if an oscillator is working just by touching the probe to its output.

Usually, the oscillator side of the injection capacitor is the best place to start. If there's a healthy signal there, indicated by a dc reading on the voltmeter, move the probe to the other side of the capacitor; it could be open you know.

Back when tube-type oscillators were more common, one way of telling if an oscillator was running was with a plain dc voltmeter touched to the grid. A dead oscillator develops no grid current, so there's no dc bias voltage on the grid. If the dc voltage is present, it's a sign the oscillator is running.

This isn't much help with transistors. Oscillator stages may run with forward bias, reverse bias, or zero bias between base and emitter. It all depends on the particular circuit and the transistor type. Bias doesn't tell you whether the oscillator is running or not.

So it's better that you get accustomed to checking for the presence of rf output from the oscillator. You'll not waste time trying to fix an oscillator stage that's working after all.

For your particular vtvm or fet vm (a vom loads down oscillators too much), buy or build a simple diode probe that rectifies rf and converts it to pure dc. Be sure it has a blocking capacitor at the input; a 470-μF 2500-volt type is good, if you build your own probe. Actually, the rf probe for almost any vtvm will do if it can be connected to your instrument. Don't try to use a scope rf probe; it's only a demodulator, and won't work.

Then go through all your own transmitter and receiver equipment. On the block diagram or schematic in the manual for each set, write down the rf voltage measured on your meter at the output of each oscillator. The dc reading you get is relative; that is, it doesn't mean there's really that many volts of rf. So you need to know what's a normal reading so you can later evaluate whether an oscillator is weak or stopped.

Be sure you're measuring the right point in the stage. The output is connected by a capacitor or a transformer to the stage that follows—a buffer amp, mixer, converter or frequency multiplier.

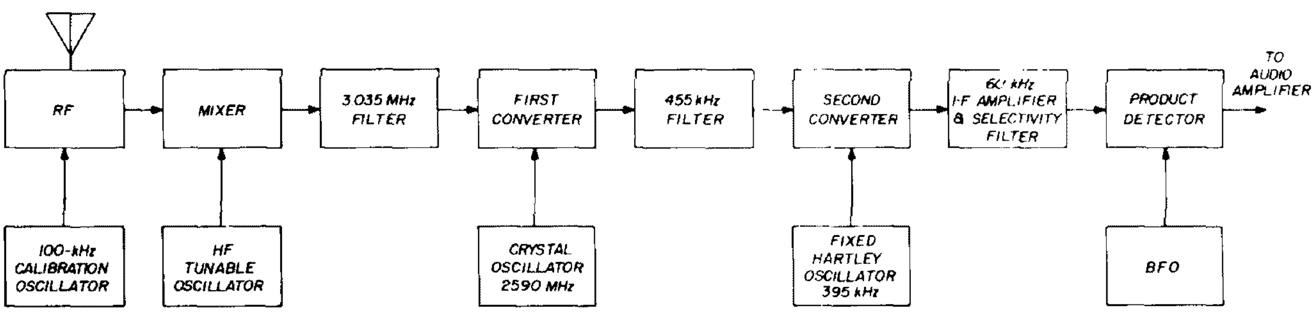



**fig. 4. Typical triple-conversation superhet uses five separate oscillators.**

## inside the stage

Once you've identified which oscillator is dead, your next job is to find what killed it. Here is a procedure that works well for me.

Take the transistor oscillator in fig. 5 as an example. It's an "inverted" version of a Hartley oscillator. Feedback to the base develops in the portion of the tapped coil between collector and B-plus. With the base grounded and the emitter not, conditions are right for regenerative feedback. The coil and capacitors in the collector resonate; the coil is variable, because this is a tuning oscillator.

Suppose you've checked at the emitter and found no rf. Start troubleshooting the stage by using your plain dc volt-

meter. Measure the regulated B-plus source. Then check dc at the collector and at the base.

At the base, the measured value is important, so check it against what the schematic or the voltage chart in the manual says is right for the set. At the collector, the value depends on too many things, particularly if there's a collector resistor. So just check to make sure the collector is getting voltage.

If the transistor is wired so the emitter gets the supply voltage, the collector should have a dc ground connection—perhaps through a transformer winding. Check that. Then make sure the emitter is getting the dc supply voltage it should.

After you're sure dc power is reaching the stage, check the action of the amplifier. One way to do this is to disable the feedback path and feed in a signal near the normal frequency of the oscillator.

In the oscillator of fig. **5**, for example, you could move the connection of C3 to the tap on the coil. Or, easier yet, just clip another .01-μF capacitor from the tap to ground. That destroys the feedback and leaves the stage operating as just an amplifier. Open C1 and you can apply an rf signal to the base. Then, with your vtvm and probe, check for rf output across R3. It should be only a little less than the signal you feed in. It also should be tunable. That is, you should get a slight change in reading as you change the frequency of the signal generator or tune the coil in the collector circuit.

In a stage like the one in fig. **1**, you just disconnect the tap between capacitors C2 and C3. Then a signal at about the frequency of the crystal, fed to the grid, should develop a signal in the cathode circuit. In the stages of fig. **2**, the crystal is the feedback device, so you just unplug (or disconnect) it. Then check amplifier operation of the tube. You can do this with virtually any oscillator stage.

Next thing to check is the feedback system itself. One way is to keep feeding the test signal in and probe for feedback energy with your rf voltmeter.

In fig. **1**, with the tap disconnected, check for a small amount of signal energy at the junction of C2-C3.

In **fig. 2**, leave only the plate end of the crystal connected. Probe at the other end of the crystal for signal energy. If you vary the generator frequency dial up and down, there should be a sharp resonant point at the crystal's frequency, revealed by a sudden sharp increase in the voltmeter reading at resonance.

In **fig. 5**, disconnect the supply end of the coil, and probe for signal energy.

Finally, you should check the oscillator for frequency. Even if it's running, an

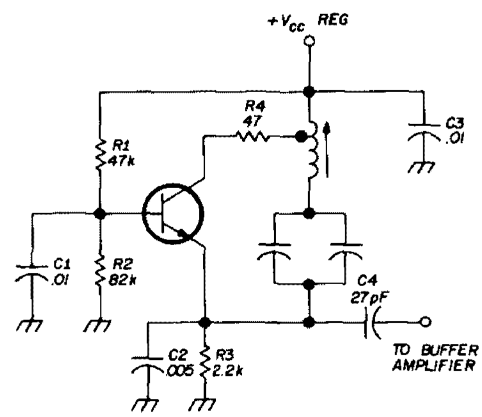



**fig. 5. Unusual version of the Hartley oscillator is inverted from the normal hookup.**

oscillator that's off frequency can't do what it's supposed to. If it's the carrier oscillator in an ssb unit, the generated sidebands that come out of the balanced modulator won't match the sideband filter. Or, an off-frequency bfo can't demodulate sidebands in the product detector.

A bad crystal is usually the cause of an off-frequency oscillator. If you know the oscillator works, yet the receiver or transmitter acts as though it doesn't, try a new crystal. If the stage is controlled by tuned circuits or a resistor-capacitor combination, check them.

You can check frequency of an oscillator with a frequency meter or with an accurate all-wave receiver set for cw re-

ception. In the latter case, a whistle or birdie shows up at the dial setting that indicates the oscillator frequency. The only trouble is, sometimes even slight crystal drift can stop operation of a set, and an all-wave receiver may not be accurate enough to spot the drift.

That sums up the way to check an oscillator. Actually, most troubles boil down to dc power or feedback. That's because you probably check the tube or transistor right off the bat. Anyway, if you check those four things in an oscillator, you'll always find the trouble without replacing whole strings of parts.

## next month

Some time ago, I promised to show you how to use a sweep generator on your repair bench. In the meantime, I've found out that some hams don't even know what a sweep generator is.

So. . .the time has come to learn. The instrument is versatile, and has many uses. I'll show you some of them.

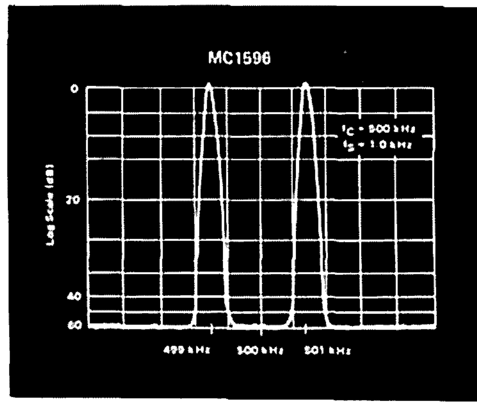
One last thing: if there's some odd circuit or set you'd like explained, drop me a line and I'll see if we can't work it in some month. After all, the purpose of this department is to help you keep your equipment working in top order. Tell us what help you need most.

ham radio


"The rig here is the latest in solid-state miniaturization except the final -- and I'm working on that."

# new products

## balanced-modulator ic

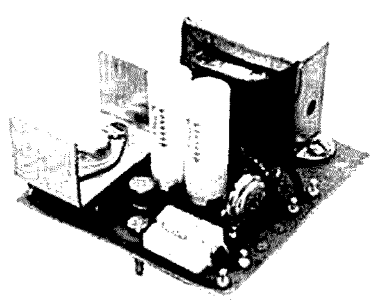



Motorola's new MC1596 balanced modulator/demodulator for communications work features typical carrier suppression of 65 dB as well as adjustable voltage gain and signal handling, balanced inputs and outputs, typical carrier feedthrough of 90 microvolts at 500 kHz and high common-mode signal rejection ratio of 85 dB typical.

The MC1596 is designed for uses where the output voltage is the product of an input voltage (modulating signal) and a switching function (carrier), such as in ssb and a-m transmission, ssb product detector, synchronous a-m detection, fm and phase detection, frequency doubling and frequency mixing. The device can be used up to 100 MHz in modulator/demodulator applications and up to 400 MHz in other circuits.

The circuit of the MC1596 integrated circuit consists essentially of an input differential amplifier driving a pair of closely matched current-mode transistor gates. These closely matched transistors are responsible for the excellent input balance that cannot be equaled by a discrete circuit of comparable cost. For more information on the versatile new MC1596, write to Technical Information Center, Motorola Semiconductor Products, Inc., Box 20924, Phoenix, Arizona 85036.

## regulated power supply



Viking Electronics has just introduced a full-range low-voltage regulated power supply for instrumentation and test applications, as well as original equipment manufacturers. The model PZ-10-A provides outputs from zero to 25 volts at currents up to 100 mA; load regulation is less than 50 mV, line regulation is ±10 mV and ripple is less than one millivolt at full rated load. The PZ-10-A has built-in provisions for remote control of current limiting and voltage output. The output terminals are floating. Also available is the model PZ-10-B which provides zero to 6 volts at 500 mA; and the model PZ-10-C which provides zero to 15 volts at 200 mA. Price of the PZ-10-A is $27.00. For more information, write to Viking Electronics of Wisconsin, 721 St. Croix Street, Hudson, Wisconsin 54016.

## hustler two-meter antennas

Two new Hustler antennas for two-meter mobile installations are optimized for maximum gain, quick installation and rugged maintenance-free operation. Both antennas are designed to handle 150 watts fm, ssb or cw, and 100 watts a-m. The heavy-duty shunt-fed base-matching inductance is completely sealed against moisture and offers maximum electrical and mechanical stability. The Hustler "break-cable" assembly provides easy coaxial feedline installation through the vehicle. The taper-ground stainless-steel radiator is field adjustable for an swr of 1.1:1 at resonance and 1.5:1 or better at 4 MHz bandwidth.

The Hustler model BBL-144 is a co-linear power gain antenna that mounts in a 3/4" hole on any flat surface; it's supplied with 17 feet of RG-58/U, a PL-259 connector and optional stainless-steel spring and 180° swivel ball. The Hustler model BBLT-144 is the same as the BBL-144 but with a Hustler trunk-lip mount for no-holes-to-drill installations. For complete specifications, write to the Sales Department, New-Tronics Corporation, 15800 Commerce Park Drive, Brookpark, Ohio 44142.

## 5-element, 20-meter beam

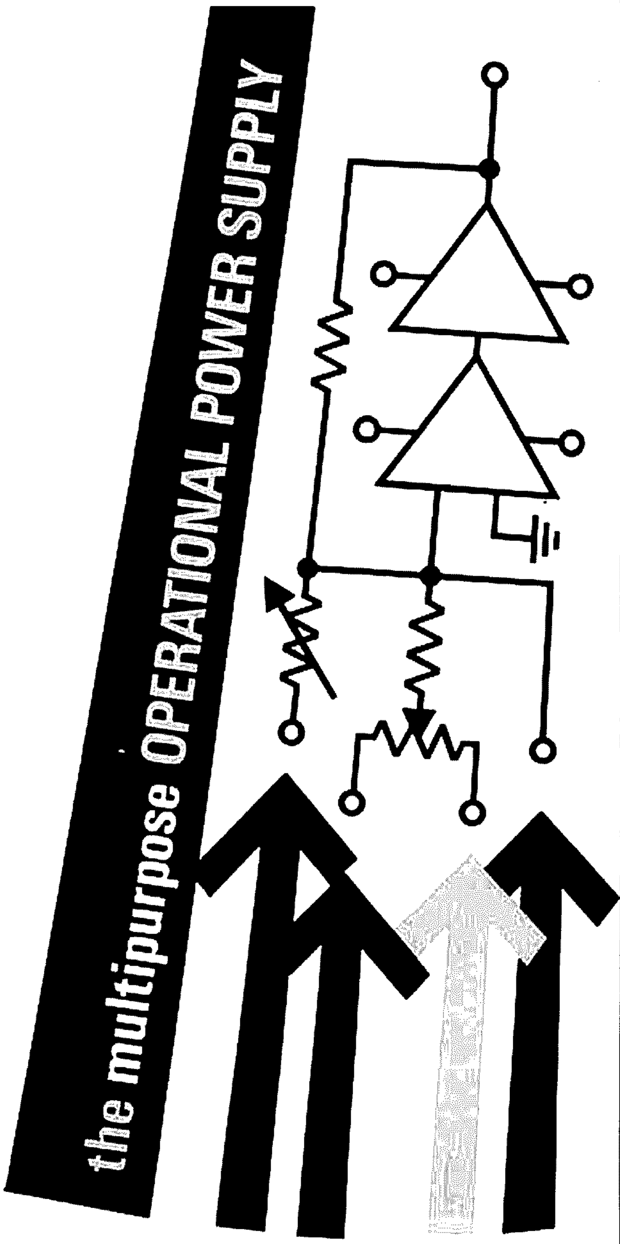
Mosley Electronics has just announced a five-element 20-meter single-band beam, the Classic 20. The unbalanced-capacitive matching system used on the Classic 20 is combined with optimum element spacing to provide maximum gain, increased bandwidth and more efficient performance. The Classic 20 antenna features high-impact insulators and clamping blocks, aluminum tubing and stainless steel hardware. The antenna is rated at 2 kW PEP ssb (1 kW cw or a-m), and exhibits an swr of 1.5:1 or less over the band. The boom is 46 feet long, and the antenna requires a turning radius of 28 feet. Assembled weight is 139 pounds. $285.51 amateur net. Mosley Electronics Inc., 4610 Lindbergh Boulevard, Bridgeton, Missouri 63042.

75 cents

focus
on
communications
technology . . .

# ham radio magazine

APRIL 1970




this month

april 1970
volume 3, number 4

**staff**

James R. Fisk, W1DTY
editor

Nicholas D. Skeer, K1PSR
vhf editor
J. Jay O'Brien, W6GDO
fm editor

A. Norman Into, Jr., W1CCZ
Alfred Wilson, W6NIF
James A. Harvey, WA6IAK
associate editors

Jean Frey
art director
Wayne T. Pierce, K3SUK
cover

T. H. Tenney, Jr. W1NLB
publisher

**offices**

Greenville, New Hampshire 03048
Telephone: 603-878-1441

**ham radio** *magazine* is published monthly by Communications Technology, Inc. Greenville, New Hampshire 03048

Subscription rates, world wide: one year, $6.00, three years, $12.00. Second class postage paid at Greenville, N. H. 03048 and at additional mailing offices

Foreign subscription agents: United Kingdom: Radio Society of Great Britain, 35 Doughty Street, London WC1, England.

All European countries: Eskil Persson, SM5CJD, Frotunagrand 1, 19400 Upplands Vasby, Sweden.



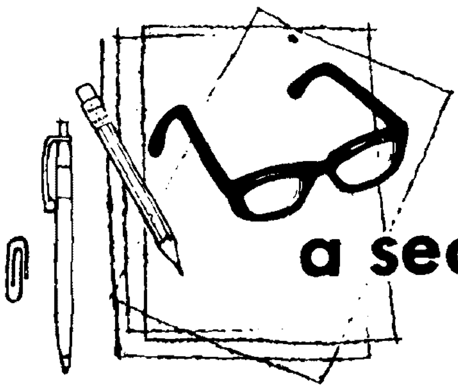
*Printed by Wellesley Press, Inc.* Wellesley, Massachusetts 02181, U.S.A.

***ham radio*** *is available to the blind and physically handicapped on magnetic tape from Science for the Blind, 221 Rock Hill Road, Bala Cynwyd, Pennsylvania 19440. Microfilm copies of current and back issues are available from University Microfilms, 313 N. First Street, Ann Arbor, Michigan 48103.

Postmaster: Please send form 3579 to **ham radio** magazine, Greenville, New Hampshire 03048.

**ham radio** *magazine*

# contents



# a second look

**As you are probably aware,** amateur radio is not in the middle of one of its more noticeable growth periods. In fact, licensing figures over the past year or so indicate that we're just about holding our own, and that's all. This may seem a bit unusual in these days of exploding population and growing technology, but when you look further and try to define the place where ham radio fits into our society it makes a little more sense.

No longer do you have to have an amateur radio license to taste the thrill of two-way radio communication. CB has opened this door to thousands, some of whom might otherwise have become hams. This unquestionably has had an adverse effect on the numerical growth of amateur radio, but perhaps it also has a bright side.

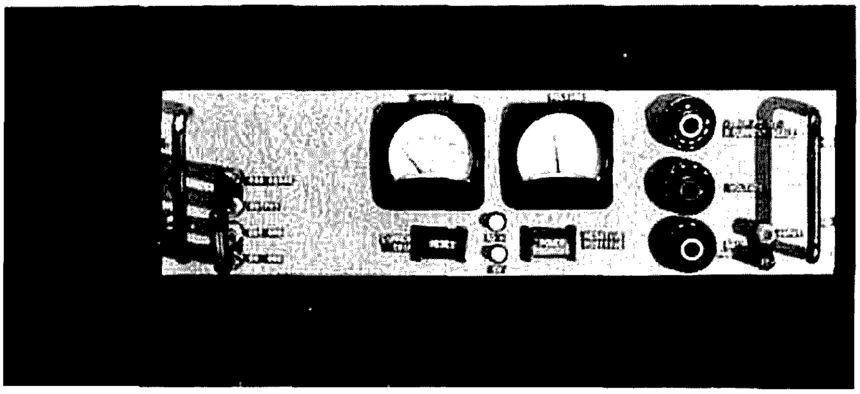
It takes a good deal of inertia to learn the requirements for a ticket and to organize one's self into a study program to prepare for the exam. After passing the test the new licensee must put together a station and actually get on the air. It doesn't just happen; he has to work for it. It has been estimated that as many as 50% of those passing the novice test never get on air!

Where does this lead us? We feel, in a very positive direction. It's always easy to achieve size by sacrificing quality; amateur radio is growing in stature and capability. For example, who would have dreamed a few years ago of the remarkably efficient and dependable local communications which have been established on 2-meter fm? The exciting accomplishments of the Oscar program and moonbounce have certainly helped make ham radio a hobby which can show much pride of achievement. Our growing mastery of RTTY, slow-scan tv and facsimile further attest to our desire to explore and develop new ideas. The transition to ssb has opened up technological and communications capabilities virtually undreamed of twenty-five years ago.

We must continually re-evaluate ourselves to make sure that we are taking maximum advantage of our privileges. We must also consider what future changes might do. We have all seen such ideas as the mighty incentive controversy and the liberalizing of novice license provisions. Currently new ideas such as putting technicians on 10 meters are being evaluated.

Some of these ideas will work out while others will be scrapped. However, we *must not* rely on the past to provide a blueprint for the future. It's important that we try to structure our activities to take advantage of any interesting new challenges which we find. Communications is in for a number of violent changes in the years ahead and amateur radio should do its best to be in the middle of the action -- satellite repeaters and data transmission techniques are just two possibilities that come to mind. It goes without saying that your shiny new ssb transceiver will be just as obsolete in the years to come as a two-tube regenerative receiver is today. *Ham radio* will do everything it can to guide you into this new age.

**Skip Tenney, W1NLB**
**publisher**



# operational power supply

**Regulated dc supply, low-frequency amplifier, current amplifier with bipolar output— these are just a few uses for this versatile instrument**

Richard Factor, WA2IKL, 30 West 60th Street, New York 10023

The amateur literature is replete with power-supply articles. They range from the simplest transformer-rectifier-filter to sophisticated units with variable voltage control and super regulation. What do they all have in common? They're all designed to operate other equipment.

It seems a shame to use all this expensive regulation circuitry for such limited purposes. This article describes a power supply so versatile that it's called a "power supply" only because there's no generic term more applicable. The primary purpose of this unit is *not* to furnish power to another instrument. It is, instead, an instrument in its own right. In addition to being a highly regulated dc-voltage supply, it also functions as a low-frequency amplifier (flat response from dc to 10 kHz) with high-power output. Or it may be used as a current amplifier. The "supply," operating both in current and voltage modes, has bipolar output that can be varied continuously through zero. Thus, an arbitrarily low output may be obtained.

To protect any load, a variable current trip disconnects the output when the current exceeds any settable value between 10 mA and 3 amps. As an added convenience, the output-voltage meter has an automatic range switch. Also featured is a circuit that provides auto-

matic polarity indication so the meters can be used on ac and dc.

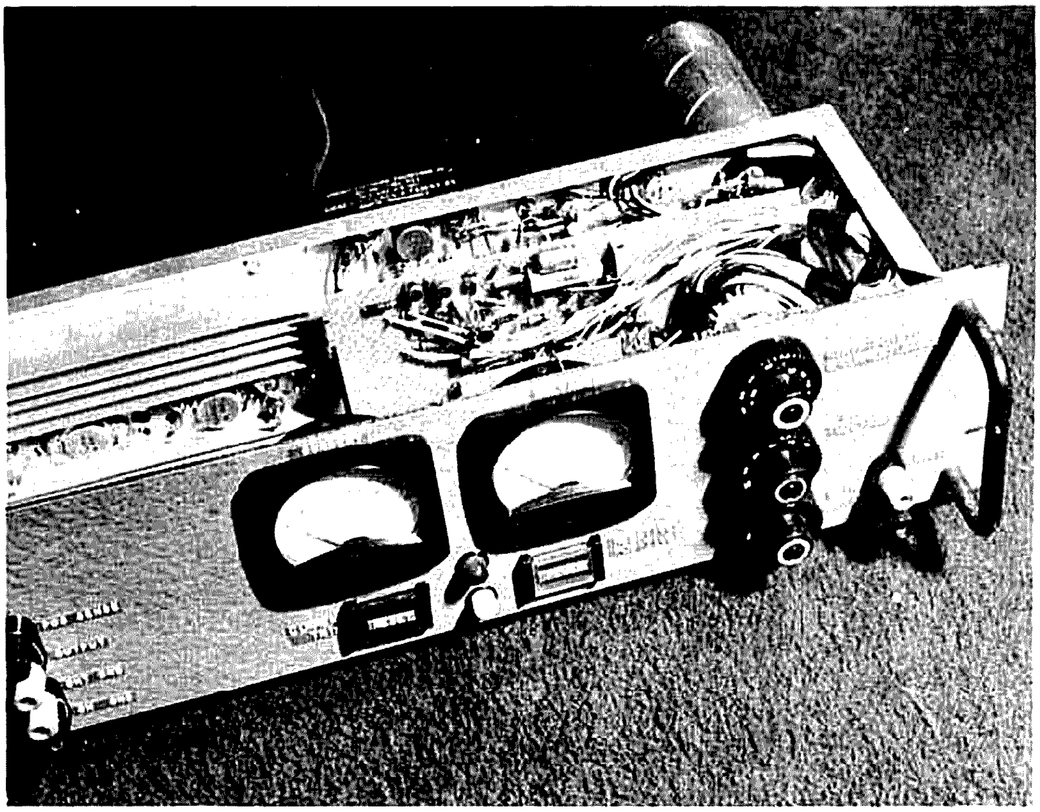
Most of the critical circuits are contained in five readily available IC's. These include two type 709 operational amplifiers, one for regulation and one for metering; and three type 710 voltage comparators, one each in the range switching, polarity indicating, and overload protection circuits.

## theory of operation

The circuit is basically an operational-

feedback around the transistor, nothing can compensate for these variations.

**Fig. 1B** can be a hundredfold better. It overcomes both of the above limitations. Very little current flows through the zener, and feedback through the zener and amplifier transistor compensates for the change in $V_{be}$. The principal objections are that the output isn't variable (except by changing the zener); also, if tight regulation is required, the gain of a single amplifier transistor isn't high enough to provide it.



The operational power supply.

amplifier-controlled series regulator modified for bipolar operation. To understand how it works, let's review the theory of the series regulator and the operational amplifier. **Fig. 1** shows several series regulators, increasing in complexity and effectiveness.

**Fig. 1A** is a "regulator" with no feedback. It is merely an emitter follower with a fixed voltage on the base. Regulation is poor because the current through the zener varies substantially, and the $V_{be}$ of the series pass transistor varies with the output current. Since there is no

**Fig. 1C** shows a differential amplifier. The circuit amplifies the difference between the reference voltage (supplied by the zener), and a sample of the output voltage (supplied by the potentiometer). This circuit allows the output voltage to be varied. Regulation percentage is again limited by the gain of the transistors.

## op amp regulators

The best regulation is obtained with circuits that use operational amplifiers. The op amp is one of the most nearly universal analog circuits. An ideal opera-

tional amplifier has infinite voltage gain and zero output impedance. Obviously such characteristics are unattainable in practice, but op amps are almost always operated with a feedback network to control gain. This makes the assumption of ideality nearly correct.

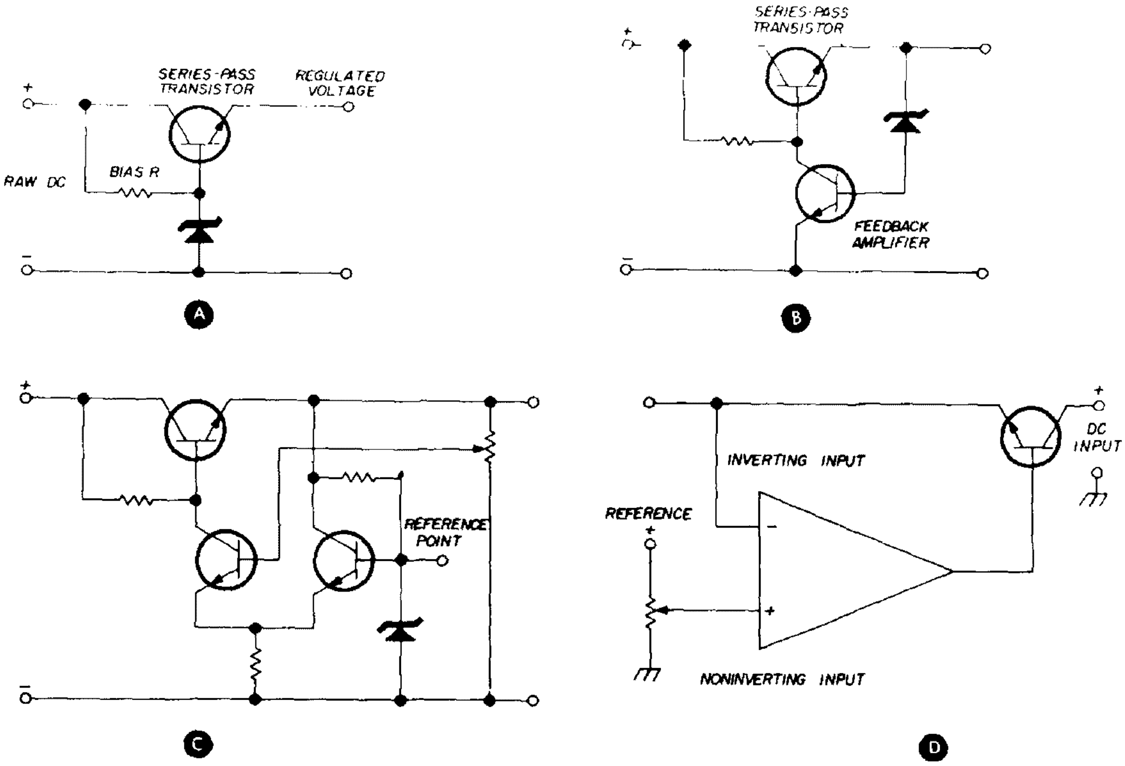
When an op amp is operated without feedback, it is said to be operating "open loop." Its open-loop chacteristics are as follows:

1. With no input voltage, output voltage is zero.

2. If the inverting input is more positive than the noninverting input, the output voltage assumes the most-negative potential of which it is capable, just slightly more positive than the negative supply voltage. This is true regardless of how much voltage is applied to the input.

3. If the noninverting input is more positive than the inverting input, the output situation above is reversed.

4. If both inputs are shorted and a voltage is applied between them and



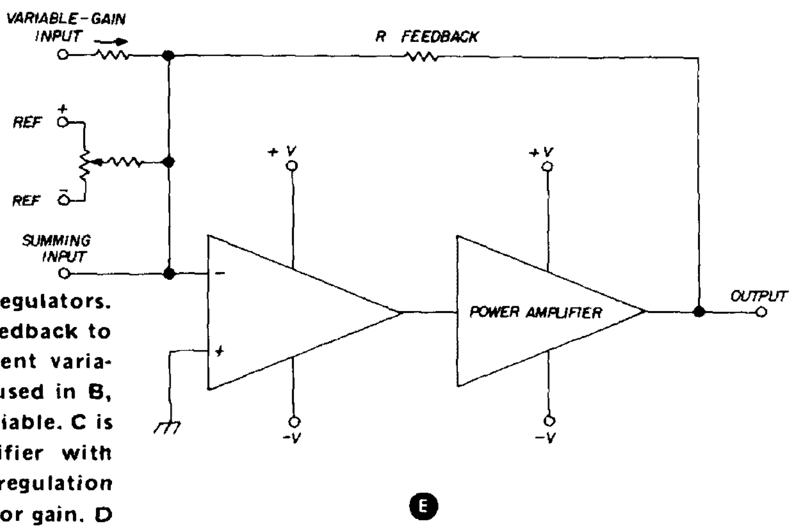




**fig. 1. Basic series regulators. Circuit at A has no feedback to compensate for current variations. Feedback is used in B, but output is not variable. C is a differential amplifier with variable output, but regulation is limited by transistor gain. D is an op amp regulator. Circuit at E is the "Operational Power Supply".**

ground, no change in output voltage occurs. This is called "common mode rejection."

5. The input impedance is infinite.

To see how the op amp operates as a regulator, consider fig. 1D. The inverting input is connected to the output of the series pass transistor, and the non-inverting input is connected to a reference voltage. If any disturbance (such as a load) is introduced into the circuit that causes the inverting input to become negative with respect to the reference, the output voltage will immediately increase to compensate for it. Thus, the output voltage will be always equal to the reference. Since the input impedance is infinite, no current will be drawn from the reference.

## input charcteristics

Now we come to the most important derived property of the op amp. When enclosed in a negative-feedback loop, its differential input voltage will be zero. As in the above example, the output voltage will always vary so the two input terminals will be at the same potential. If the noninverting input is grounded, the inverting input will become a virtual ground, and any input voltages connected there (through resistors) will become pure currents. Several inputs can be connected simultaneously without any interaction.

Fig. 1E is a block diagram of the circuit used in the unit. Let's examine the properties of the inverting amplifier when a resistor (Rfeedback) is included in the loop. Recalling that the input impedance is infinite, it's obvious that the voltage drop across the resistor is zero. Thus, there is just as much feedback without the resistor as with it. This is not precisely true, but true enough for a very good approximation.

Let Rfeedback be equal to 100 kilohms. So far we have a circuit whose output voltage is zero, since the non-inverting (reference) input is grounded. Connect a 10 kilohm resistor to the inverting input, along with the feedback resistor already there. Put ±1 volt on the other end of this resistor. The inverting input is at zero potential, so 100 μA must flow through the resistor. The op amp's input impedance is "infinite," so the current can't go there. To balance the current and the voltage, the output voltage must be -10 volts, so that -100 μA must also flow through the feedback resistor to balance the circuit. By using these resistors, we have an inverting amplifier (also called a summing amplifier).

The application of this circuit to a power supply is obvious. Take a positive and negative reference, connect a variable

Rear view of the operational power supply showing the power transistors and adjusting controls.

resistor between them with its wiper connected to the input, and we have an extremely stable, bipolar, variable-voltage source. If a power amplifier is connected inside the feedback loop so that it actually becomes *part* of the op amp, we will have what I'll call an "Operational Power Supply," or *OPS*.

## power supply or amplifier

I've avoided mentioning anything

other than dc voltages so far. Take a look at figs. 1A through 1E. None of the circuits depend on capacitors for filtering or regulation. This is important because it greatly enhances the time response of the supply. The output will change from fully positive to fully negative in 100 μs. Since it is so fast, the operational *power supply* could just as easily be used as an *amplifier.*

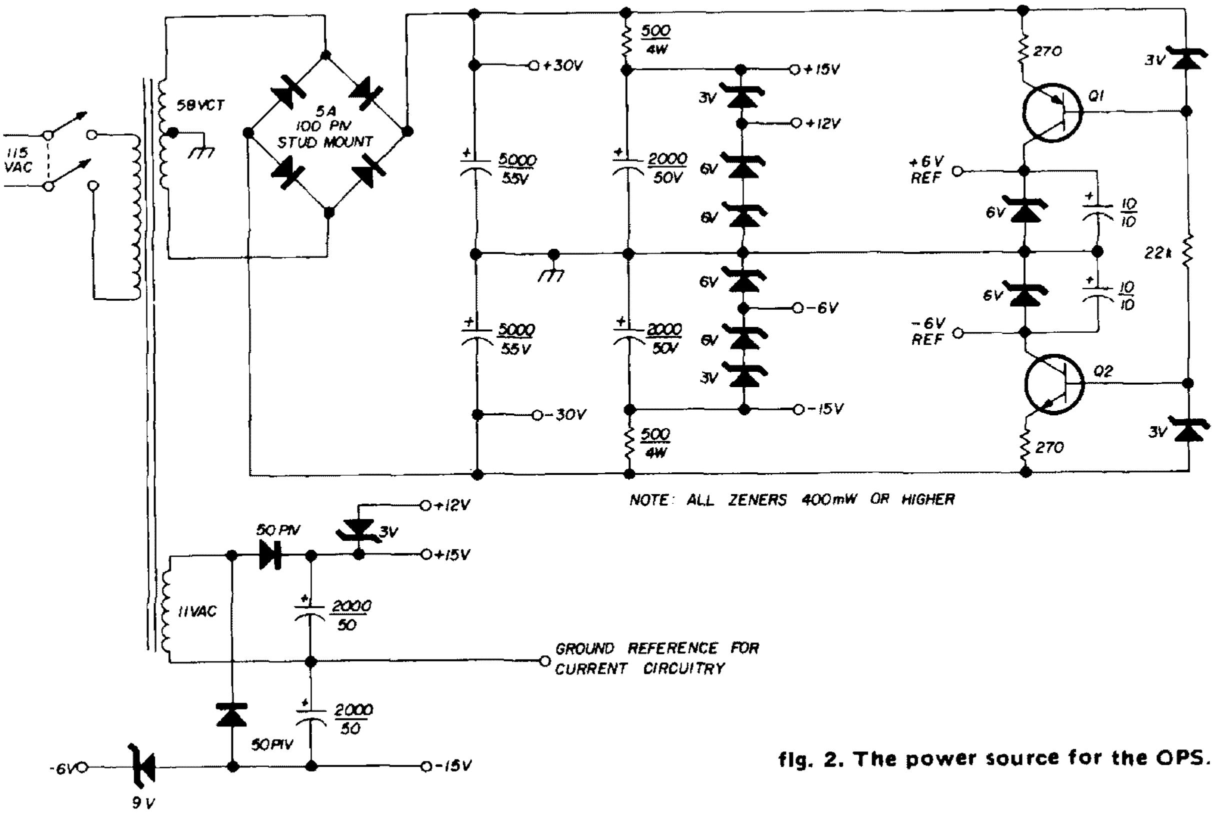


fig. 2. The power source for the OPS.

## construction

The photos show the circuit layout and panel-component arrangement I used. However, these can be varied to suit your requirements and available parts. Most of the components were on hand. A wide range of values should be acceptable. *Probably the biggest problem you'll have* is finding a commercially available transformer that has a 58-volt ct secondary at 3 amps and an 11-volt secondary at 100 mA. By connecting filament transformer *windings in series, it should be possible to* obtain voltages sufficiently close to those shown.*

*A Variac in the primary might be helpful to make up whatever input voltages are necessary. Editor.

The IC's are available from several manufacturers. Any silicon diode with low leakage is okay for the signal diodes. Output transistors are HEP247 (npn) and HEP248 (pnp). Any silcon transistor with a 5-ampere or higher rating may be substituted. $V_{ce}$ and $V_{cb}$ should be at least 60 volts; 80 volts would be even better.

All zeners should have 1-watt dissipation or more. Their voltage values are shown in the schematics. All other transistors are 2N2905A (pnp) and 2N2219A (npn). Any device with similar ratings may be used.

Feel free to make substitutions. However, don't use electrolytic capacitors with values much lower than those specified. Critical components, such as meter shunts, are discussed in the parts of the article that describe the circuits. Be sure to use heat sinks for the two output transistors and rectifiers (photo).

## power source

Fig. 2 shows the basic power source for the entire unit. Several voltages are required for the output stage, the IC's, and the reference supply. These are:

| | |
|---|---|
| +15<br>+12<br>-6<br>-15 | Referenced to ground for the regulator and voltage-metering portions. |
| +15<br>+12<br>-6<br>-15 | *Referenced to the output bus for* the current metering and overload-protection circuitry. |
| +6<br>-6 | Stable, referenced to ground, to drive the operational amplifier input. |

The first two groups are obtained by simple resistor-zener divider chains. The last group is obtained by using a positive and negative current source to drive the zener. Q1 and Q2 each have six-volt zeners in their collector circuits, with a fixed voltage drop across their emitter resistors. Since zeners are voltage stable with respect to current changes (except for a small slope after the "knee"), regulating the current keeps the zener operating over a very small portion of its curve, with a proportionally smaller change in voltage.

## regulator circuit

**Fig. 3** shows the actual circuit represented by **fig. 1E.** Theoretical operation has already been explained. Circuit details follow.

The two diodes connected back-to-back from the inverting input to ground protect the amplifier from high-voltage input transients. The capacitor and resistor between pins 1 and 8, and the capacitor between pins 5 and 6 give the proper high-frequency gain roll-off characteristics to the op amp.

The 100 kilohm resistor connected to the input is the main feedback resistor for the circuit. It is connected to the "sense" terminal on the front panel, which is normally shorted to the power-amplifier output.

Note that, in this circuit only, the inverting and noninverting inputs are reversed as to function, so don't be disturbed if the pin numbering doesn't correspond to that in the other diagrams. The reason for the reversal is that the power amplifier is itself an inverting amplifier, although one with relatively low voltage gain.

The circuit has one stage of voltage gain (Q3 and Q4), which drives compound emitter followers (npn Q5 and Q7; pnp Q6 and Q8). The diodes and 250-ohm potentiometer between the collectors of Q3 and Q4 are a biasing network that eliminates crossover distortion at output voltages near zero. The 250-ohm potentiometer should be adjusted for 60 mA quiescent current

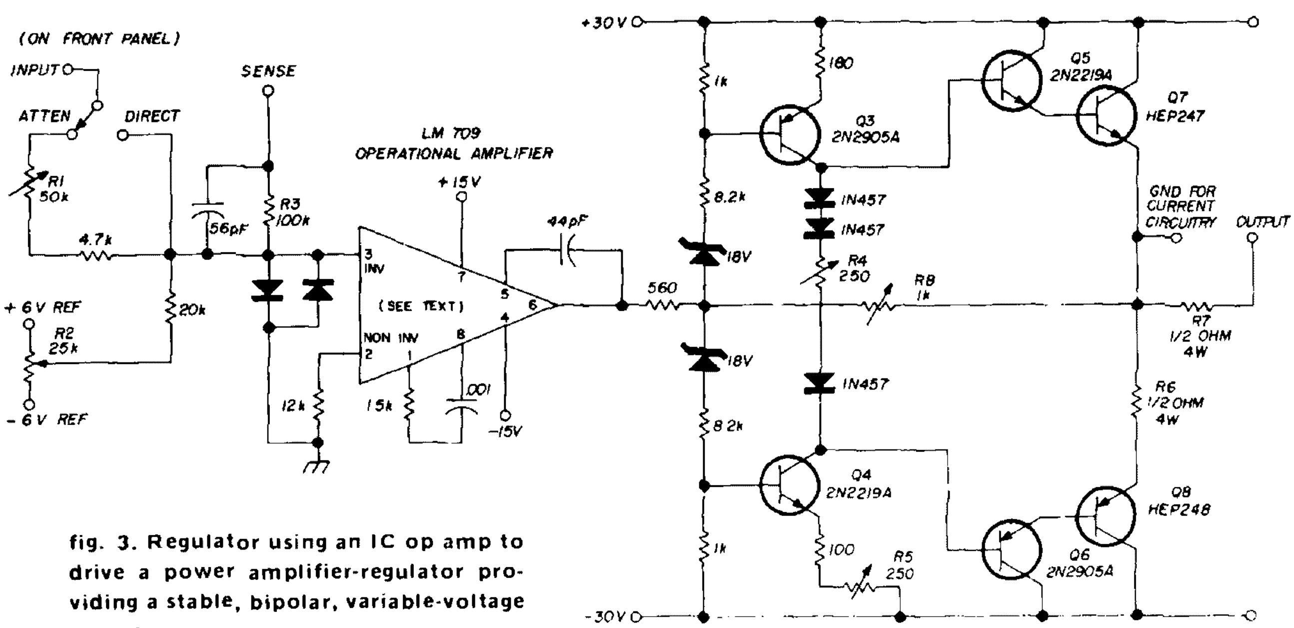



**fig. 3. Regulator using an IC op amp to drive a power amplifier-regulator providing a stable, bipolar, variable-voltage source.**

through R6, by measuring 30 mV across it. R5 *balances out differences* in the bias network of the voltage amplifier. It is adjusted by setting the supply output voltage to zero with R2, then measuring the voltage at pin 6 of the 709. The correct adjustment is reached when this voltage is between ±1 volt. R8 should be adjusted so that at least a 16-volt peak-to-peak excursion (measured again at pin 6) produces a 60-volt peak-to-peak excursion at the output terminal. R7 is a shunt for current metering and is duplicated in **fig. 4**.

## current metering

The usual method of measuring power supply current is to connect a meter in series with the supply. However, with highly regulated supplies, the output impedance is much lower than that of the meter, and the meter's insertion will seriously degrade performance.

The circuit of **fig. 4** eliminates this difficulty. The ½-ohm resistor in series with the output is connected inside the feedback loop,.so any voltage drop across it is compensated. A milliammeter of arbitrary value (less than 1 mA full-scale to avoid overloading the op amp) is connected in a bridge rectifier in the 709's feedback loop. Different series resistors are then switched to the inverting input.

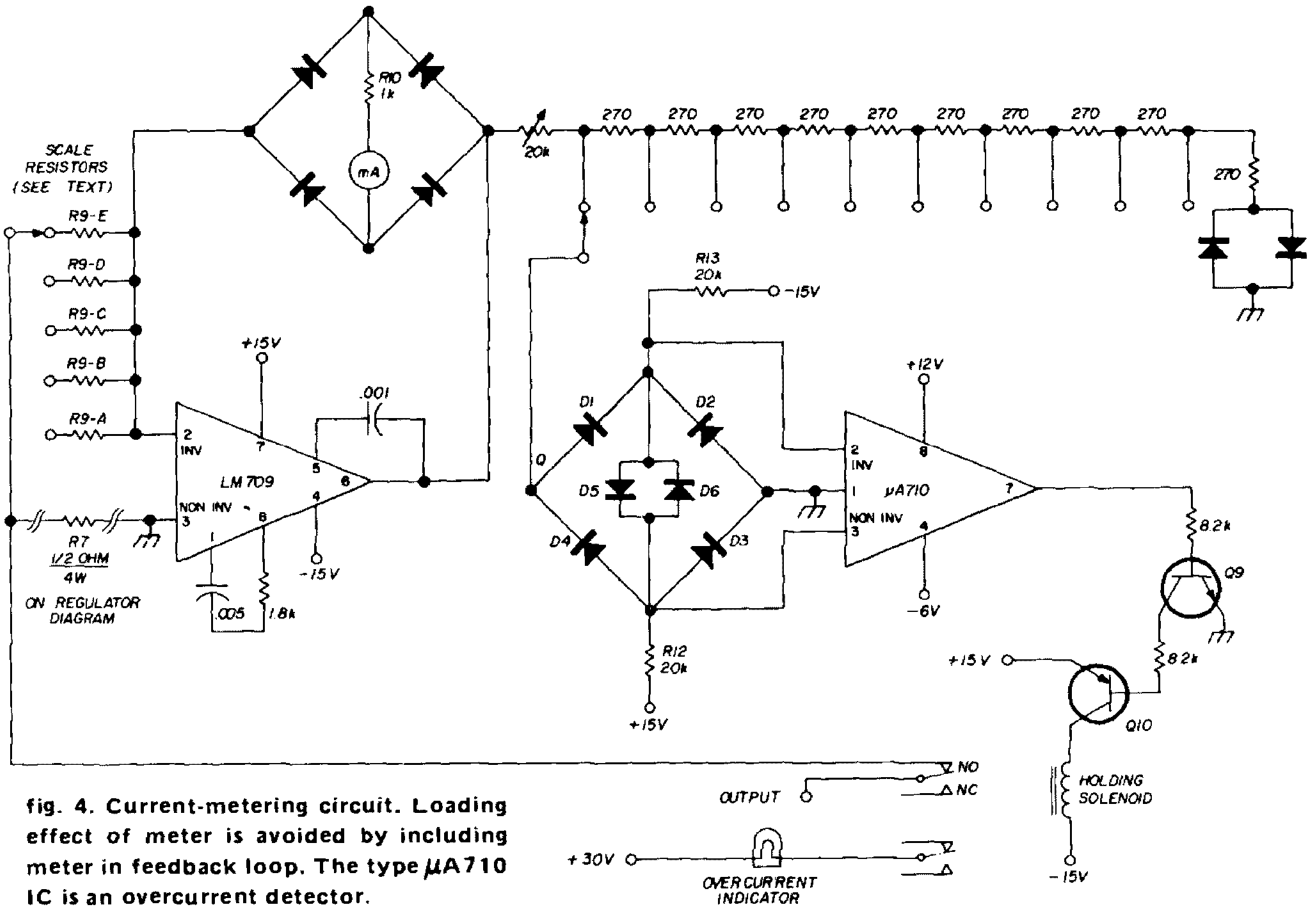



**fig. 4. Current-metering circuit. Loading effect of meter is avoided by including meter in feedback loop. The type µA710 IC is an overcurrent detector.**

Since the op amp input voltage is always zero, calculating the resistor values is very simple. Assuming the current you want to measure is 50 mA or greater and the meter is 1 mA (or less), just calculate the voltage across the ½-ohm resistor that corresponds to the full-scale current. *Then find the resistance* across which this voltage will give 1 mA. The characteristics of the meter do not enter into the computation. If you want to measure very low currents, it's necessary to cal-

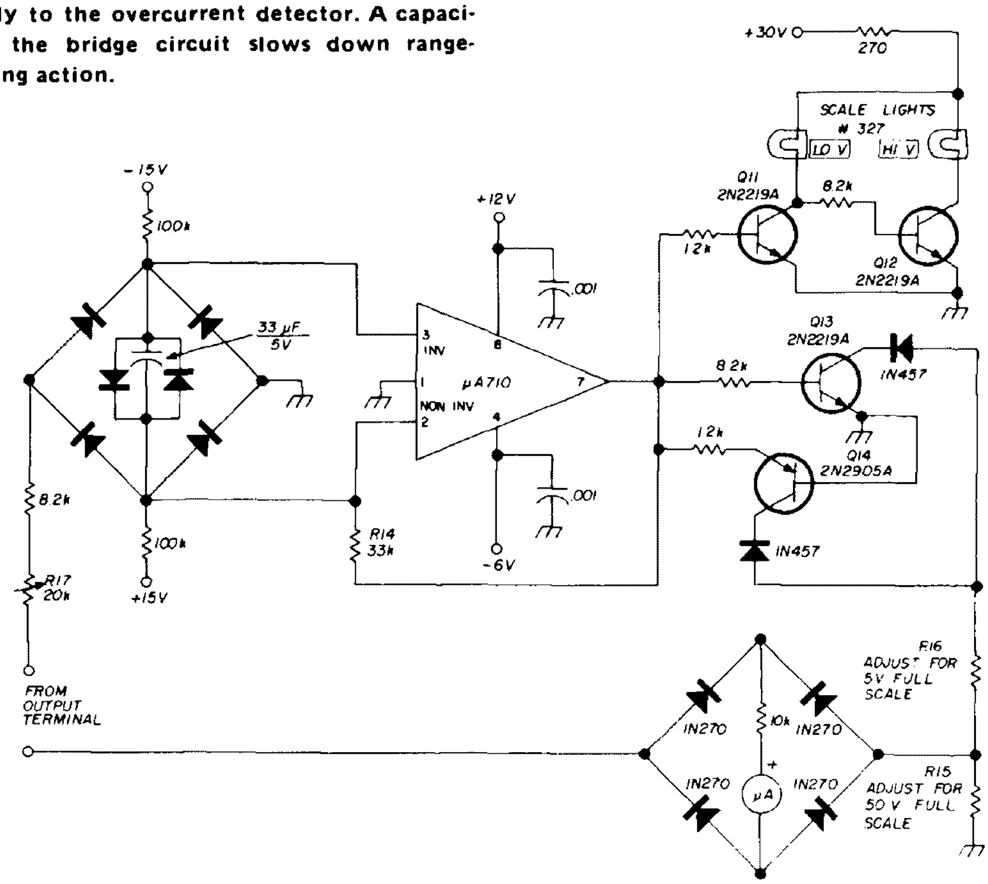
culate the shunting effect of the metering resistor across the ½-ohm resistor, R7.

## overcurrent detector

A type 710 voltage comparator is used as an overcurrent detector (**fig. 4**). It is similar to an op amp; however, instead of having a linearly varying output, its output voltage is either close to zero, if the inverting input is positive with respect to the noninverting input, or about 3 volts when the situation is reversed. By setting a reference voltage on one input, the comparator gives a digital output indicating whether the other input is higher or lower. The comparator is biased by R12 and R13 so that the output is at 3 volts. This biases Q9 into conduction as well as Q10, causing current to flow through the coil labeled "holding solenoid." The voltage at point Q is derived from the output of the op amp, which is in turn determined by the supply output current and the range-switch setting. R11 is adjusted so that, with the divider switch in its least-sensitive position, the voltage at point Q overcomes the fixed bias on the comparator input at a point just after the current meter goes off scale. Changing taps on the divider enables cutoff to be set at several discrete points withing the meter range.

The output interruptor is part of a switch that has a light and a solenoid. Current through the solenoid is sufficient to hold the switch in, but not to pull it in. Thus when the current is interrupted as the trip point is reached, the switch pops out and lights up, and must be manually reset. It would be just as simple to use a surplus 24-volt relay and reset button here, if you have no surplus computer switches handy. D1-D4 enable the comparator to work on ac, and D5-D6 protect the inputs from excessive voltage.

**fig. 5. Voltage-metering. The $\mu$A710 IC operates similarly to the overcurrent detector. A capacitor in the bridge circuit slows down range-switching action.**



## voltage metering

After all the foregoing complexity, it might disappoint you to learn that the voltmeter is merely a simple voltmeter. The range switching circuit (fig. 5) is a bit unusual, however. The comparator in this circuit works in a manner similar to that in the current trip. The only difference is that a 33 $\mu$F capacitor is connected across the input to slow down the range switching action. The 33 kilohm resistor, R14, provides some hysteresis so that the up-range and down-range switch points aren't identical. Q11 and Q12 illuminate the appropriate scale lights so that you know what you are reading, and Q13 and Q14 shunt the low-range resistor to ground for both positive and negative excursions of the output.

## range resistor adjustment

The range resistors are adjusted as follows. Disconnect the wire going to the shunting transistors, and set the supply output for 30 volts. Adjust R15 so the meter reads 30. Set the supply output for 5 volts. Adjust R17 so the high-range light just comes on. Reduce the output voltage to 4.5 volts. If the low-scale light hasn't come on yet, increase the value of R14 until it does. Now reconnect the wire to Q13 and Q14 and adjust R17 for a reading of 4.5 on the meter. Readings will be somewhat inaccurate below 2 volts, because it was necessary to use silicon diodes in series with the shunt to reduce leakage.

For the voltmeter, I used a 0-50 microammeter. It makes little difference from a voltage-metering standpoint what you use, since the meter puts an insignificant load on the supply. However, as noted previously the supply can be used in a constant-current mode, in which any load in parallel with the desired one reduces the current regulation of the supply. If you don't have a very sensitive

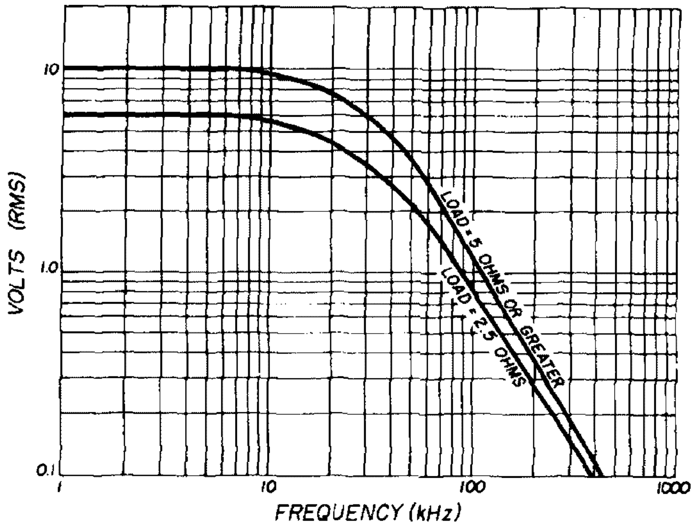
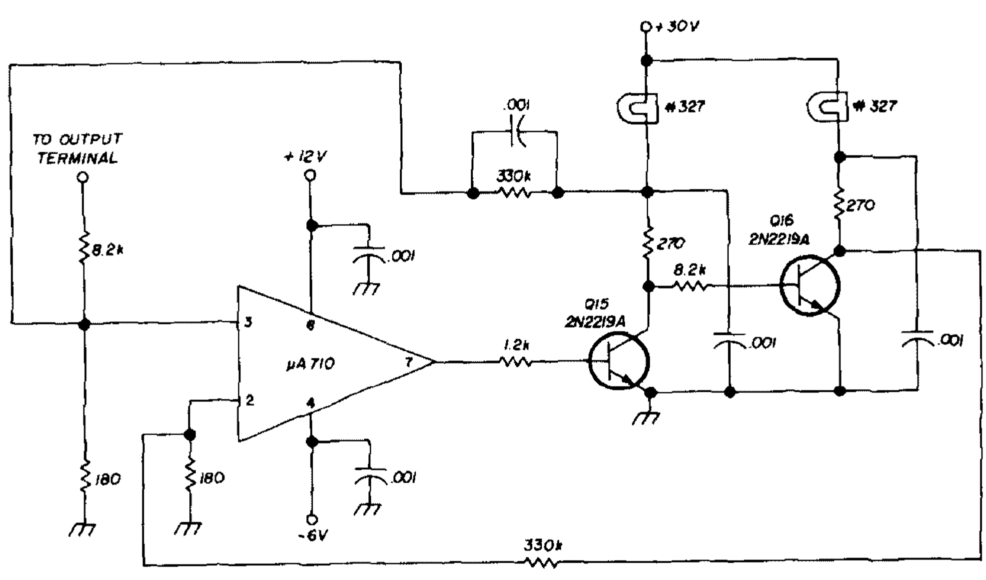



**fig. 7. Frequency response of the OPS. Slope above 10 kHz is 20-dB/decade roll off to prevent oscillation.**

**fig. 6. Positive-negative output indicator. The IC comparator indicates +dc, -dc, or ac output.**



meter, it's possible to use an operational amplifier to boost meter sensitivity. The bibliography at the end of the article gives several methods of doing this.

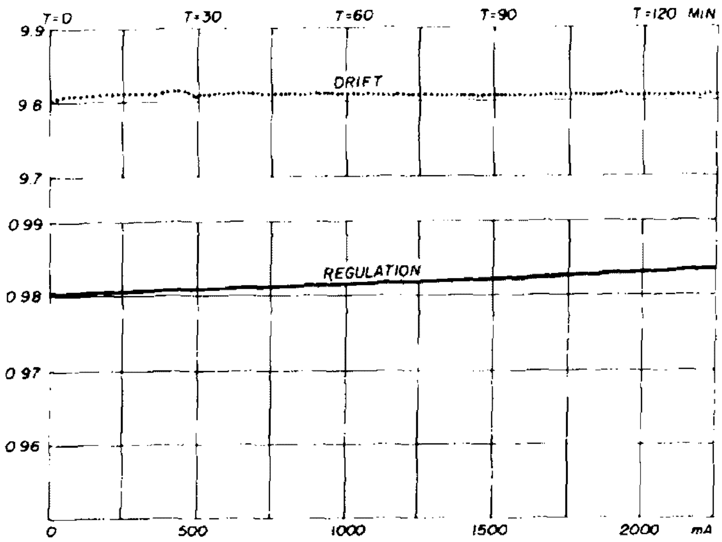
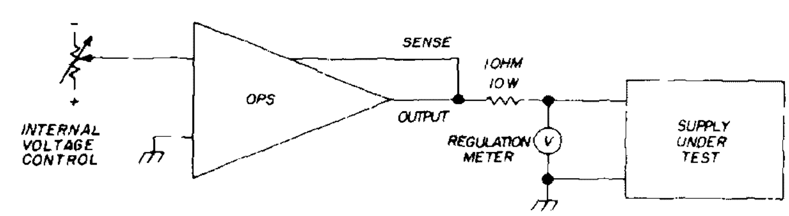



**fig. 8. OPS regulation and drift performance.**

## positive-negative detector

There are several reasons for using a positive-negative output indicator **(fig. 6)** instead of a zero-center meter. One is that zero-center meters are difficult to come by. Another is that by using the entire scale to read voltage or current, resolution is improved 100 percent. Also, when measuring ac, a zero-center meter will indicate only the dc component of the signal. By arranging the metering so that both a positive and a negative output give

**fig. 9. The OPS as an electronic load. By varying input voltage, loads can be simulated without resistors.**




positive meter deflections, no switching is necessary for proper readings. The comparator circuit indicates if the output is *+dc, -dc or ac (if both lights are lit).*

The comparator works by using ground as a reference instead of a bias. Thus, if a voltage is below ground, one light is on; if above, the other is illuminated. The capacitors prevent the extremely fast operation of the comparator (under 50 nanoseconds) from radiating noise to the rest of the circuit.

## performance

The drop-off in frequency response **(fig. 7)** might seem rather sharp, but only a portion of the graph is shown. The remaining response is flat down to dc. Above 10 kHz, a 20-dB-per-decade roll off is required so that the amplifier gain drops as the phase shift increases.

The regulation curve **(fig. 8)** may seem somewhat anomalous since it shows an increasing output voltage as the load increases. Actually, nothing particularly strange is happening here. The regulation at the terminals of the op amp is normal, but the distribution of the lead resistance inside the unit makes it appear otherwise. There are techniques for compensating for such effects, but there's little point in making the regulation better than the random variations of the output.

## applications

So now you've built this fascinating gadget, with some blinking lights. You can't put it on a Christmas tree, nor is it heavy enough to make a decent boat anchor. So what to do with it? You can use it as a combination that has characteristics neither an amplifier nor a power supply alone possesses. The only limit is your ingenuity. Below is a compilation of a few ideas I came up with.

## electronic load

Due to the bipolar nature of the OPS,

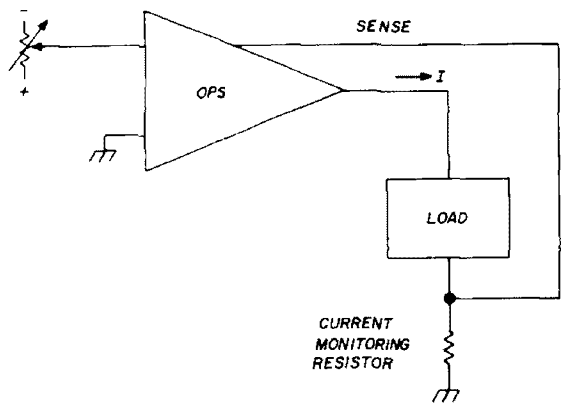



**fig. 10. Current regulator application. Load is part of the feedback loop; a change in load resistance causes a corresponding output voltage change to maintain constant current.**

it can absorb current as well as produce it. With a 1-ohm, 10-watt resistor, your power resistor problems are over. Assume you're testing a 10-volt power supply. Connect the resistor as shown in **fig. 9.** By varying the voltage on the ops from 10 to 7 volts, the load current changes from 0 to 3 amps. To test the supply with load resistors would require several, or at least one, with a 30-watt rating.

## current regulator

The supply can be used as a current regulator. All that's required is a small change in the output connection **(fig. 10)**. To maintain a constant voltage across the current-monitoring resistor requires a constant current through it. The voltage, and hence the current, are determined by the dc-input current and the value of the current-monitoring resistor. The load is part of the feedback network, so a resistance change of the load causes a corresponding output voltage change to maintain constant current. If a capacitive

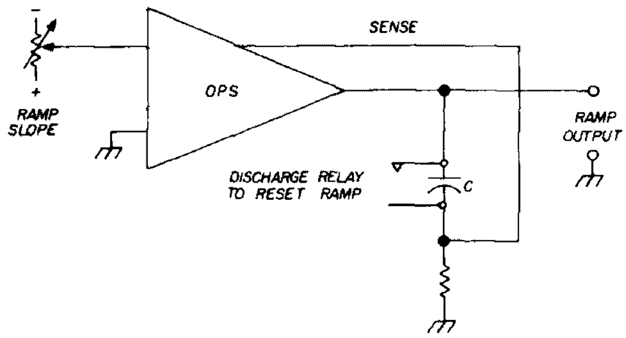
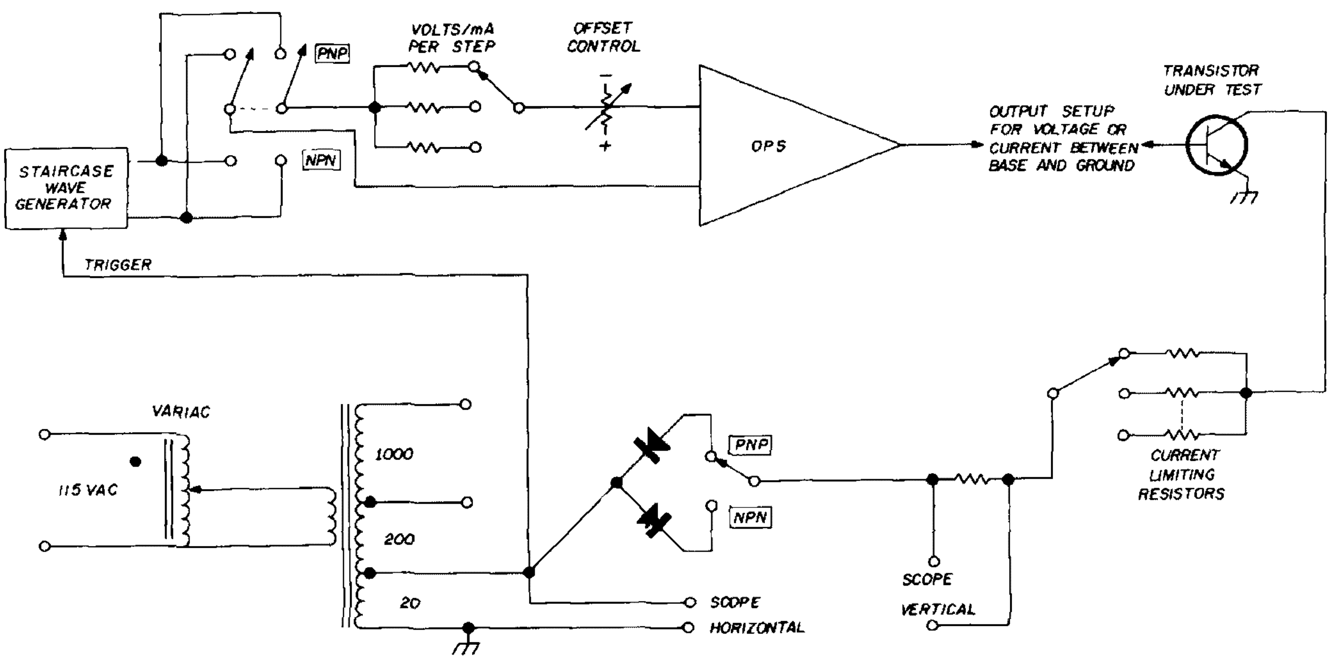



**fig. 11. A linear ramp generator results if the capacitive load, C, is discharged automatically.**

load is used, the capacitor will be charged with a constant current, giving a linear increase in output voltage. By providing a method of automatically discharging the

**fig. 12. Semiconductor curve tracer. Bipolar output of OPS allows it to be used with npn or pnp transistors.**



capacitor, a linear ramp generator is created, **fig. 11**.

## semiconductor curve tracer

An application where constant current, constant voltage, and programmability are necessary is in a curve tracer **(fig. 12)**. Obtaining a single curve is possible by ordinary methods, but it's more useful to have a family of curves at different base currents or voltages. Due to the bipolar nature of the supply, it can be used for npn and pnp resistors, including power transistors.

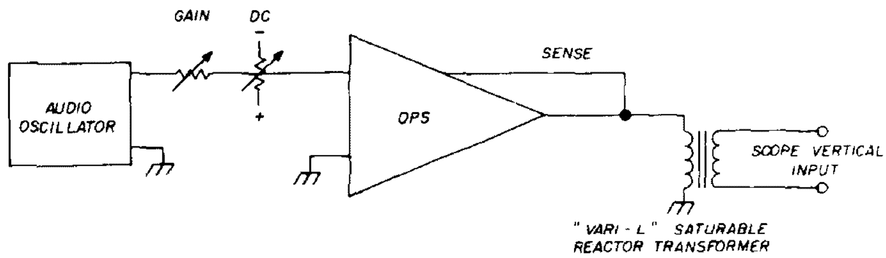



**fig. 13. Inductor or transformer analizer. Core saturation is indicated when the secondary voltage decreases or becomes nonlinear.**

## inductor-transformer analyzer

Other testing applications might include determining characteristics of audio inductors or saturable reactors **(fig. 13)**. Apply a small ac voltage to the input of the supply, connect the output to a transformer primary, and slowly increase the dc output. When the output voltage at the secondary decreases or becomes nonlinear, the core is saturating. This is a good way to find out the current rating of modulation transformers. This hookup can also be used to sweep variable inductors[1], while providing a constant bias.

## temperature-controlled oscillator oven

Are you trying to build an extremely stable crystal oscillator? The most difficult parameter to control is the crystal temperature. Mechanical thermostats typically regulate only within a degree or two. Put a thermistor in the crystal oven **(fig. 14)** and connect it in a bridge arrangement so that the bridge is nulled when the thermistor is at the proper temperature. Connect the OPS to detect the null voltage. Connect the output directly to the oven winding (disable the thermostat), and you will have a proportionally controlled oven with a temperature stability at least an order of magnitude better than a thermostat.

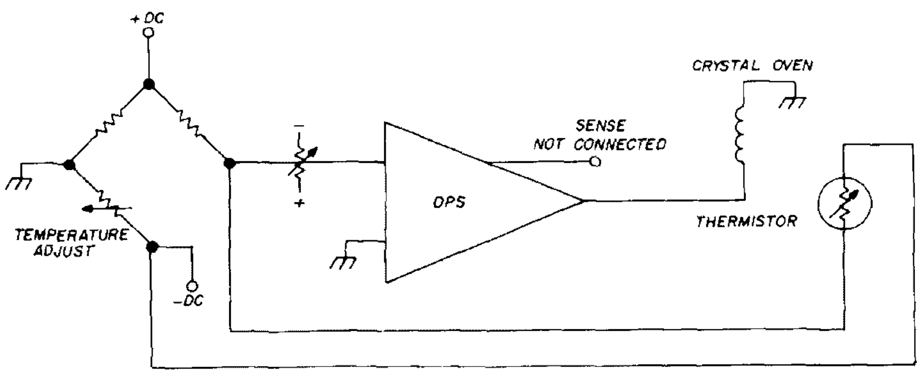



**fig. 14. Proportional temperature controller. The ops provides better control than a thermostat.**

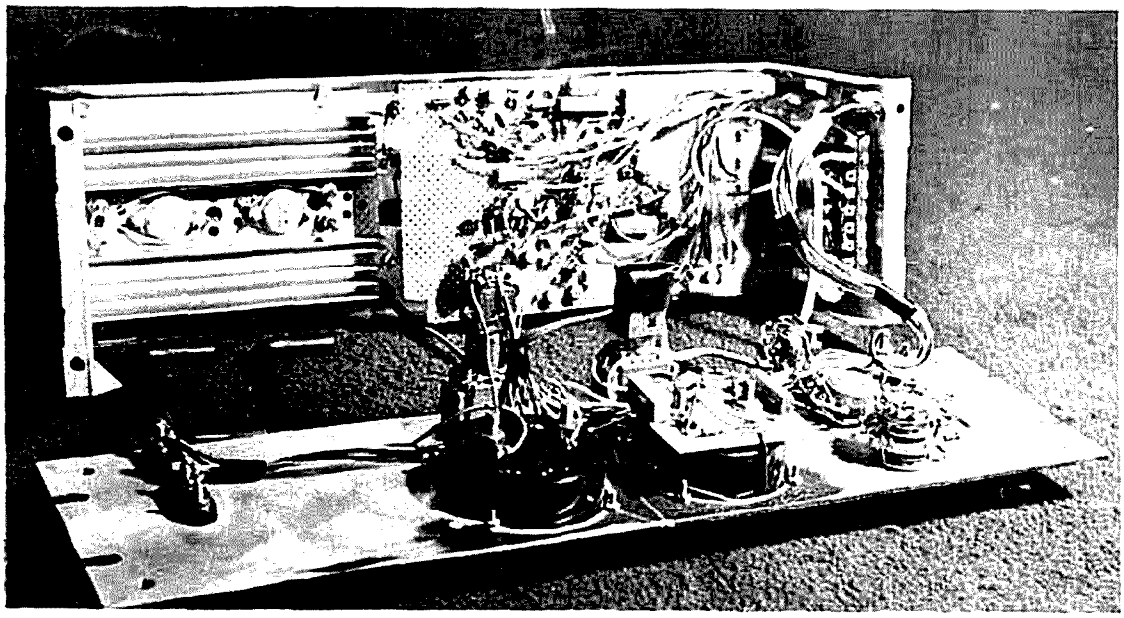
## tracking power supply

Power supplies of equal positive and negative voltage are often required for IC experimentation. Since the operational power supply is an inverting amplifier, it can be adjusted to give a gain of -1. Connected to the output of an ordinary supply, it will give an equal and opposite output voltage that will track the voltage of the original supply. ***Fig.* 15** *is an* example.

## conclusion

After you've built your OPS, you'll be quite familiar with two of the most popular IC's on the market. You'll also be able to boast to your hi-fi-nut neighbor that you have an amplifier with better bass response than his.

I'd like to acknowledge the help of Steve Schwartz, WA2YDN, who made the photographs accompanying this article.



Inside view of the OPS. The perforated board behind the voltage meter contains the meter multipliers. The heat sink hold the two output transistors and the rectifiers.

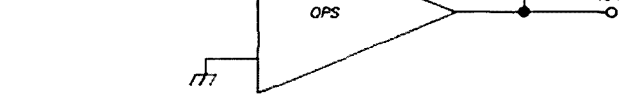



fig. 15. Tracking power supply. An equal and opposite output voltage from the OPS will follow that of the power supply being tested.

### reference

[1] Robert M. Brown, "Solid-state Current Controlled Tuning", *ham radio*, January, 1969, p. 38.

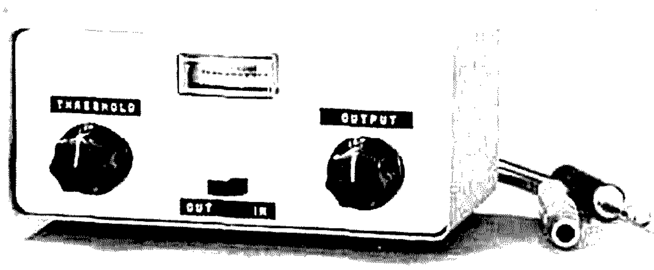
### bibliography

*Handbook of Operational Amplifier Applications*, Burr-Brown, Inc.
*Handbook of Operational Amplifier Active RC Networks*, Burr-Brown, Inc.
*Linear Integrated Circuits Applications Handbook*, Fairchild Semiconductor, Inc.
*DC Power Supply Handbook*, Application Note 90, Hewlett-Packard.
*Power Supply Handboook*, KEPCO, Inc.
*Operational Amplifiers*, Radiation, Inc.

ham radio

## next month, a special antenna issue featuring:

triangular beam antennas
multiband dipoles
80-meter beam
antenna matching techniques
integrated swr/power meter
antenna tuner design
receiving antennas
the isotropic antenna
swr alarm circuits
cubical quads




# simple speech processor for ssb

R. Lee Price, K6PHT, 1512 Oberlin Avenue, Thousand Oaks, California

Anyone who listens critically to amateur ssb stations will notice many signals with poor audio quality. Even though the transmitters putting out these signals are well designed, distortion will occur if transmitter design limitations are exceeded. Such distortion is caused by attempting to exceed peak available power or maximum average power.

This article explores methods for obtaining maximum transmitter effectiveness, while maintaining basic power limitations.

## modulation control

In many commercial stations, volume compressors are used between the microphone and transmitter. The compressors are of two types. One operates as an average program level control. It has relatively long time constants to maintain output at optimum average modulation level. The other, operating as a peak limiting device, has shorter time constants to prevents over-modulation on peaks.

Optimum average modulation control and peak level control also apply to amateur transmitters. Regardless of peak envelope power or the type of compressor, limiter or alc circuit used, all ssb transmitters are power limited by output tube plate dissipation and, of course, the power supply's capability and regulation.

If an ssb transmitter is rated at, say, 500 watts peak envelope power, then with no speech processing circuit the ratio of peak-to-average power would be of the order of 5:1 for normal speech. Average power input would be about 100 watts. With an efficiency of 60 percent, about 40 watts would be dissipated in the final amplifier tubes. If these tubes have a rated plate dissipation of more than 40 watts, then some increase in average power would be permissible if it could be obtained without increasing peak envelope power. This can be done with a speech processor that controls the ratio of peak-to-average power in the audio signal.

## volume compressors

A volume compressor is a form of automatic gain control. Agc is obtained by rectifying a portion of an audio amplifier's output; the rectified output then controls audio gain when fed back to circuits preceding the audio amplifier. Volume range is thus automatically controlled as a function of input signal level.

In a volume compressor, volume range

is automatically reduced, and the average power level of the speech signal is increased relative to its peak. In a peak-limited ssb system, such a circuit can be used to increase average modulation without a corresponding increase in peak modulation level.

Because the speech envelope rises and falls, the rate of the speech envelope variation is referred to as the "syllabic rate." This occurs between about 0.5 and

agc that will increase average power level and readability through noise, before the accompanying wave-form change (distortion) becomes objectionable, depends on received signal-to-noise ratio.

## peak clippers

A second type of circuit for increasing average modulation level is the peak clipper or "hard limiter." This has very short attack and decay time constants.

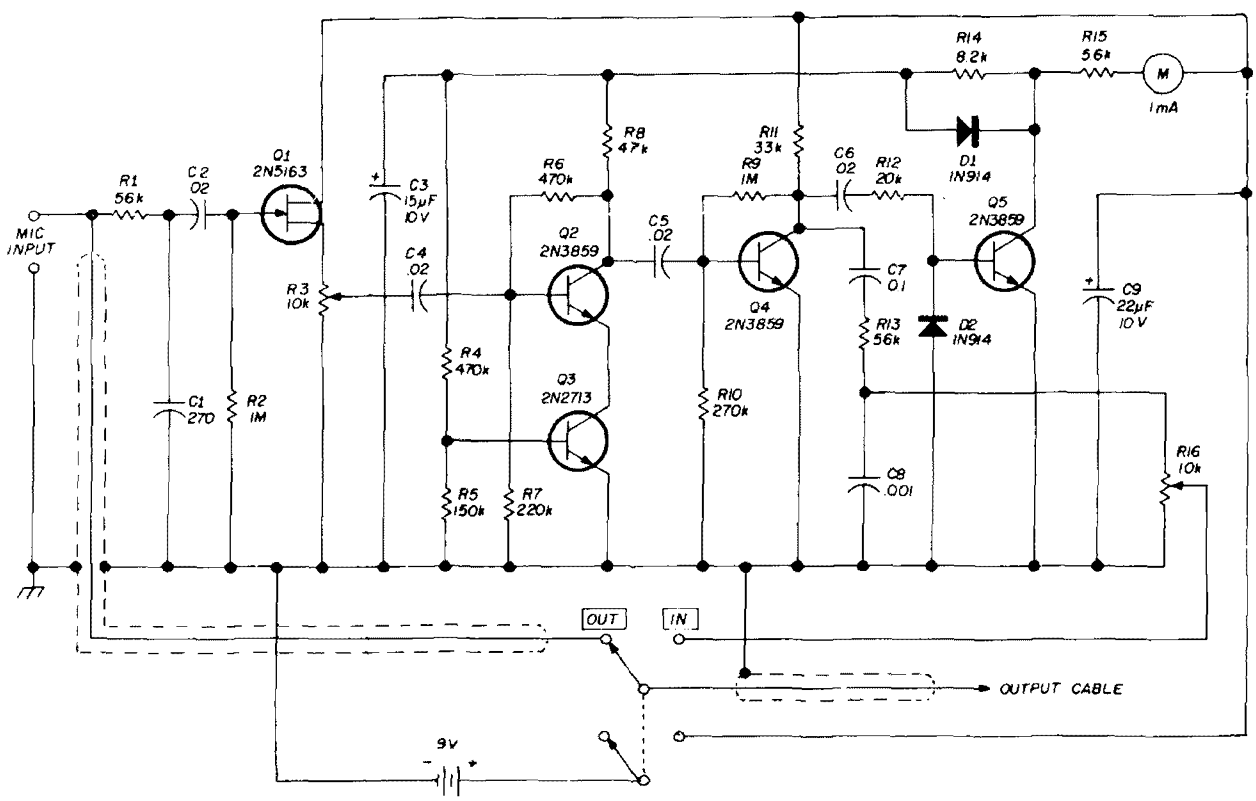



fig. 1. Schematic of battery-operated agc audio signal processor. Distortion is less than 3 percent on a 1-kHz sine wave with 14 dB compression.

25 Hz, with an average of about 5 Hz. The speech envelope variation is, in effect, a form of amplitude modulation. The amount of a-m is of the order of 25:1, or 28 dB for average continuous speech. Much of this a-m can be removed without greatly effecting speech intelligibility. However, if too much of the syllabic rate variations are removed, low-level syllables and breath sounds become excessive and speech has an unnatural sound.

An oscilloscope will show that compression causes a change in speech waveform, which is interpreted by the ear as distortion. The amount of compression or

When used at audio frequencies, distortion is introduced because the audio waveform peaks flatten. Although most of the distortion can be removed with low-pass filters, we are left with intermodulation distortion products within the audio passband. This distortion tends to reduce intelligibility.

Peak clippers or limiters have been used effectively at radio and intermediate frequencies. If carrier frequencies are properly chosen, harmonic and intermodulation distortion products, resulting from flattening of the rf signal peaks, will fall outside the rf or i-f passband and can be removed with filters.

## alc circuits

The alc circuits included in most ssb transmitters are compressors operating at the output radio frequency. Single-sideband transmitters use frequency converters to translate speech frequencies to a band of radio frequencies. After translation, the ssb signal is linearly amplified. Examination of the ssb rf signal shows a variation in amplitude at the syllabic rate not unlike the original speech waveform envelope. The alc circuit has a compression threshold near maximum modulation to prevent flat-topping. Because of this high threshold, alc circuits don't have a dynamic range of more than a few dB, which isn't sufficient to compensate for large variations in audio signal level.

## agc speech compressor

An effective speech processor is an agc *amplifier inserted in the microphone* circuit. Not only does the processor maintain optimum modulation level; it is particularly useful when the transmitter is being modulated by someone other than an experienced operator. In phone patch work, for example, phone line signal levels vary between persons and from one connection to the next. An agc speech processor between phone patch and transmitter will equalize audio signal level variations.

After much experimentation, I developed the agc process shown in **fig. 1.** Its features include:

1. Simplicity and low cost.

2. Battery operation. Battery life should exceed 200 hours.

3. Low distortion: less than 3 percent on a 1-kHz sine wave with approximately 14 dB compression.

## circuit description

Q1, an fet connected as a source follower, provides high input impedance. This input circuit allows almost any type of microphone to be used. Threshold control R3 provides proper signal level to the agc amplifier despite output from *different microphones.*

Q2's gain is controlled by a combination of two effects. As signal level increases above agc threshold, agc rectifier/amplifier Q5 starts to conduct. Its collector current decreases the voltage on agc capacitor C3, which decreases Q2's collector current. Q2's gain is proportional to collector current, so amplifier *gain is reduced.* At the same time, the collector impedance of Q3, which is the emitter degeneration for Q2, increases since it is also biased from C3; thus Q3's bias is decreased. These two effects, decreasing collector current and increasing emitter impedance, provide a net

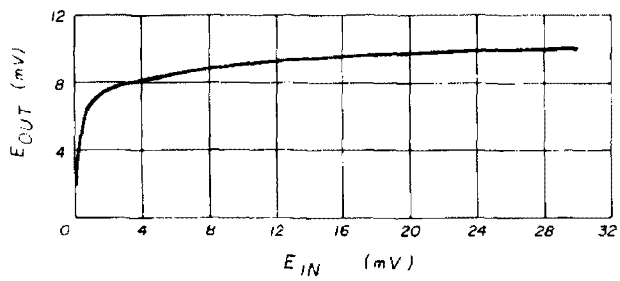



**fig. 2. Agc response with speech processor threshold set at about 2 mV.**

gain reduction without an appreciable change in Q2's collector voltage. Thus no transient "thump" is noticeable in the amplifier's output.

Q4 is a linear, constant-gain output amplifier that also provides proper threshold voltage for Q5. R12 isolates Q4 from Q5 so that Q4's output waveform won't be distorted from nonlinear loading by Q5. R1, C1 make up an rf filter for the microphone input, and R13, C8 constitute a low-pass filter for the output signal.

## specifications

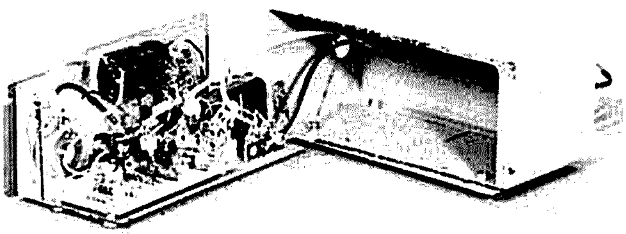
The performance of this circuit has been carefully measured with the following results:

| | |
|---|---|
| frequency response (Hz) | 300-3000 |
| maximum compression (dB) | 26 |
| attack time (ms) | 10 |
| decay time (ms) | 100 |
| distortion (percent) | 3* |

*On 1 kHz sine wave at 14 dB compression.

## operation

The threshold and output-level controls, R3 and R16, should be set as follows. Set the function switch to the OUT position, and set transmitter gain controls for proper modulation level. Set the function switch to the IN position and set output control R16 to minimum. Advance threshold control R3 until meter M1 kicks up to about half scale while speaking into the microphone in a normal voice. Turn up R16 until full modulation is again obtained. Further adjustment of transmitter gain control shouldn't be necessary with the agc amplifier either IN or OUT. A good rule for adjusting the amount of agc action by setting the threshold control is to use no more agc compression than necessary to obtain reliable communication. If signal-to-noise ratio is high, an excessive amount of compression will make speech sound unnatural. On the other hand, if signal-to-noise ratio is low, additional agc compression will raise the average modulation level.



Speech processor showing parts layout.

## use of panel meter

An advantage of this circuit is that the amount of agc in use is always visible on the meter. The threshold control may be adjusted for existing conditions by watching the meter. If signal-to-noise ratio is poor, an increase in threshold and agc level will provide better intelligibility. A low threshold setting is advisable under strong signal conditions. The plot of fig. 2 shows agc action with the threshold set at about two millivolts.

## transmitter output limitations

The object of speech processing circuits is to raise average power level. Therefore it's important that your transmitter and its power supplies operate at the resulting higher average power level. If your transmitter is limited by peak power level, then a speech processor might improve your signal's readability by increasing average power level. If your transmitter's output is limited by average plate dissipation of the final tubes, power supply, or both, there's little advantage in attempting to increase the ratio of average-to-peak power.

## other considerations

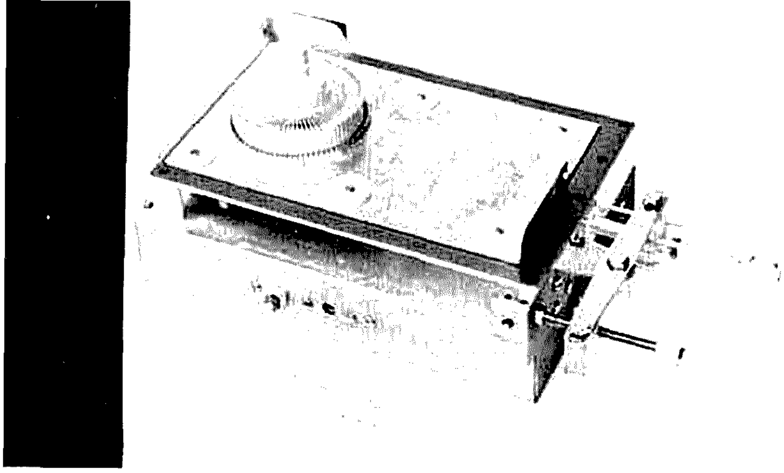
Many of the ssb transceivers on the market use TV-type sweep tubes as rf power amplifiers. It is well to remember that, although these tubes may be capable of quite high *peak* power, their *average* plate dissipation may be quite low. This limitation should be carefully considered before expecting improvement from speech processing circuits. Also, good air circulation is essential to proper operation of these power amplifiers. An efficient cooling system, with adequate space around the amplifier, is good insurance against overheating.

In conclusion, I'd like to re-emphasize that improvement in intelligibility can be obtained with speech processors when properly used, but rf power amplifier tube plate dissipation is still a limiting factor.

### bibliography


1. H. G. Collins, "Ordinary and Processed Speech in SSB Application," *QST*, January, 1969, pp. 17-22.
2. "Radio Communication Handbook," RSGB, 4th edition, pp. 9.25-9.30.
3. "Radio Handbook," Editors & Engineers, 17th edition, pp. 297-304.
4. Keith Henney, "Radio Engineering Handbook," McGraw-Hill, New York, 1959, pp. 19-72, 19-73.


ham radio



# two-kilowatt linear for two meters

A high-performance stripline amplifier tailored for the Henry Radio 2K power supply

Bob Sutherland, W6UOV, Ray Rinaudo, W6ZO, Merle Parten, K6DC

A previous article in *ham radio* described a 2-kW stripline amplifier for 150 MHz.[1] It was designed to conduct proof tests on the new Eimac 3CX1000A7 high-mu ceramic triode. As pointed out in the article, the amplifier can be adapted to two-meter operation.

We decided that this amplifier, built into the Henry 2K cabinet, would make a neat and compact two-meter package. With this in mind we obtained a 2K power supply, amplifier cabinet, and a set of panel meters.* The amplifier's plate line and output coupling system were modified, and metering and control circuits were added to accommodate the 2K supply. The photo of the completed unit shows the result: a full-legal-power linear that performs as well as it looks.

This article contains information to allow duplication of the amplifier by the serious vhf enthusiast.

## design

A schematic of the amplifier appears in fig. 1. The tube operates in a conventional grounded-grid circuit. Plate and grid current are measured in the filament return leads. An 18-volt zener in the filament return sets the desired idling plate current. A reflectometer circuit is included to aid in tuneup.

The amplifier is neutralized by moving

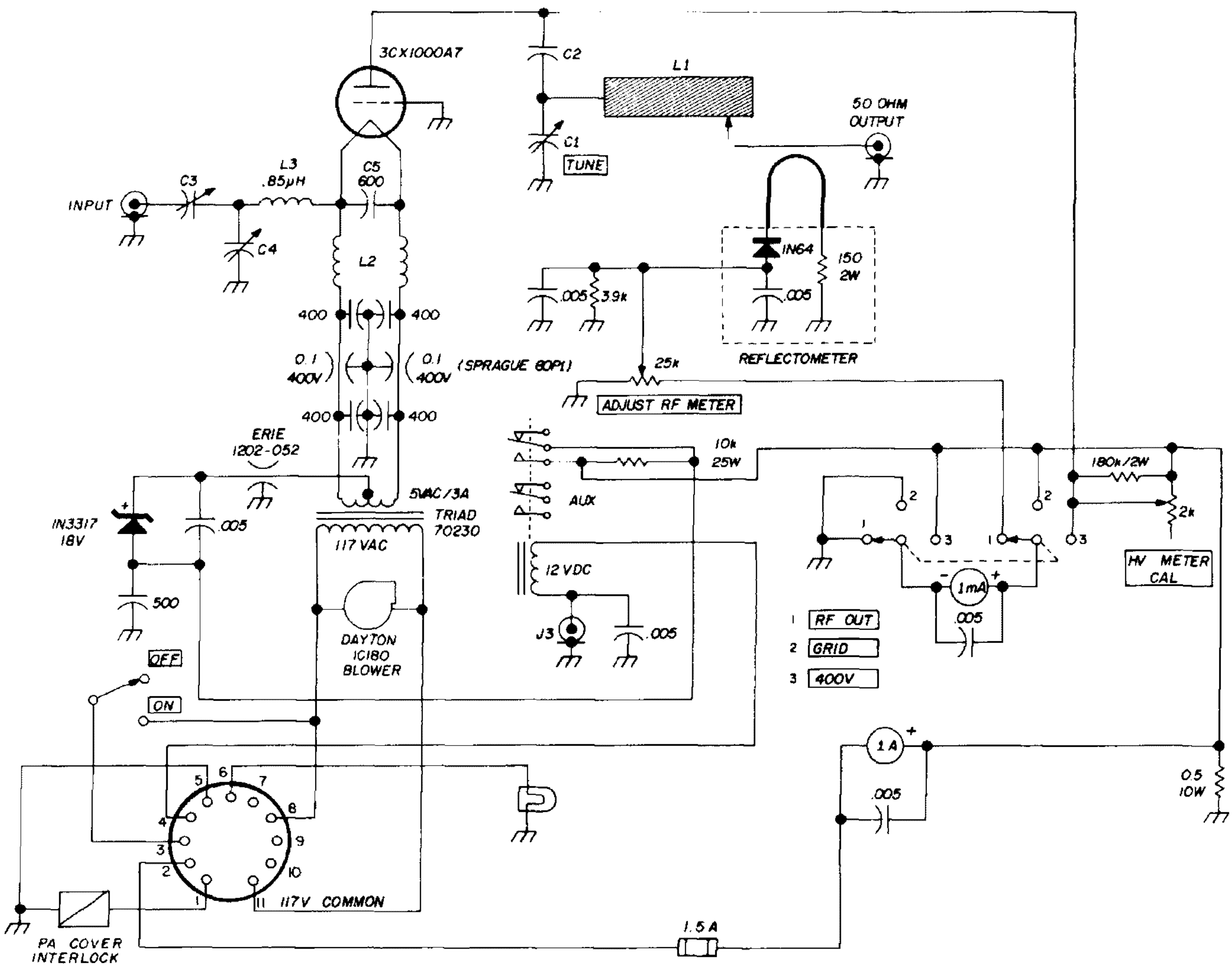
*Henry Radio Stores, 11240 West Olympic Blvd., Los Angeles, California 90064.

the self-neutralizing frequency of tube and socket from 105 to 145 MHz. The control-grid lead inductance is reduced by modifying the tube socket (see photo). The 1/8-inch cylindrical spacers and 1/16-inch rectangular spacers are re-

socket stack; then standard nuts are used.*

## cathode circuit

A T-network matches the 50-ohm drive line to the input impedance of the 3CX1000A7, which is 42 ohms.




| | | | |
|---|---|---|---|
| **C1, C2** | **Part of strip line assembly (see detail drawing available from Eimac)** | **C5** | **600 pF; six 100-pF capacitors in parallel.** |
| **C3** | **35 pF piston, set to about 3.8 pF (Johanson type 803)** | **L1** | **Part of strip-line assembly (see detail drawing available from Eimac)** |
| **C4** | **Approximately 6.5 pF. 1" diameter plates on lead screw (set to about 1.4 pF)** | **L2** | **9 turns no. 8 enameled, 3/4" ID, bifilar wound 4 turns no. 14, 3/8" ID.** |

**fig. 1. Complete circuit diagram of the kilowatt amplifier for 144 MHz.**

moved. The latter spacers have threaded holes for the long machine screws that hold the socket stack together. To secure the stack in the modified socket, the long screws are reversed and inserted into the chimney mounting holes, chassis, and

The modified Eimac SK-870 socket is mounted directly on to the chassis deck,

*This assembly, plus that of the entire plate-circuit resonator and output coupler are detailed in a scaled drawing available from Eimac, 301 Industrial Way, San Carlos, California 94070.

putting the control grid at dc ground. The grid current is metered in the cathode return lead. The photo of the chassis underside shows the filament choke and input matching network.

## plate circuit

Note the two sheet-metal plates mounted on the right end of the tuned circuit in the photo showing the top view of the plate resonator. These plates are part of the tuning capacitor, C1; one plate is movable from the front panel by an eccentric drive. A padding capacitor is mounted at the open end of the plate line. The mechanism for changing the loading (at left in the photo) is discussed later.

A metal trough runs along the cabinet wall and into the area between the front panel and the subpanel to provide shielding for metering and control wires.

The photo of the completed amplifier shows the eccentric drive, made of Rexolite 1422, which moves one plate of tuning capacitor C1. The capacitor plate is made of beryllium copper so it will return to its original position as the eccentric drive is turned. A lip on one end of the movable plate bears against the tuning eccentric.

The shoulder washers that secure the upper plate-line sandwich of copper and Isomica were machined from Teflon rod stock. The Isomica was obtained from Minnesota Mining and Manufacturing Company. Probably other dielectrics such as Teflon and Mylar could be used but Isomica has a dielectric constant between 4.5 and 5 – about twice that of Teflon; Mylar has a dielectric constant of about 3.1.

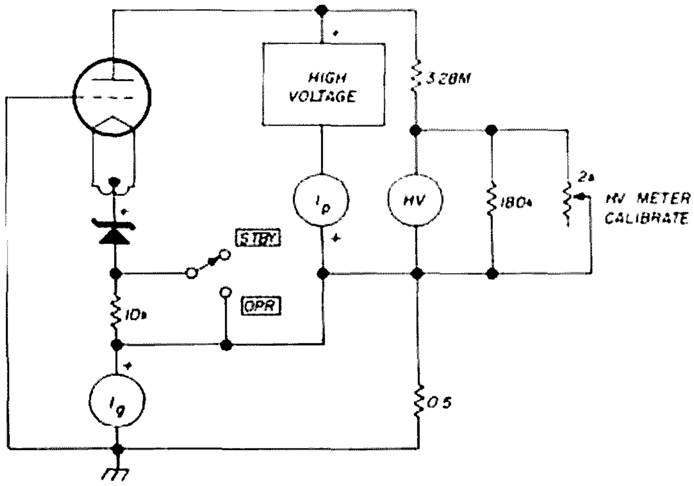
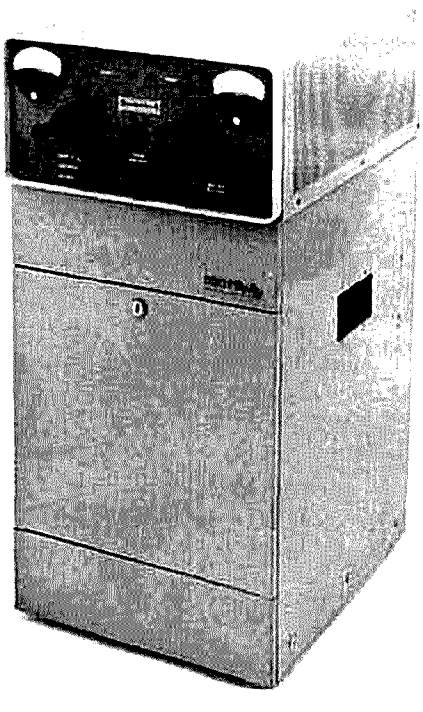


fig. 2. Basic metering circuit used in the amplifier.

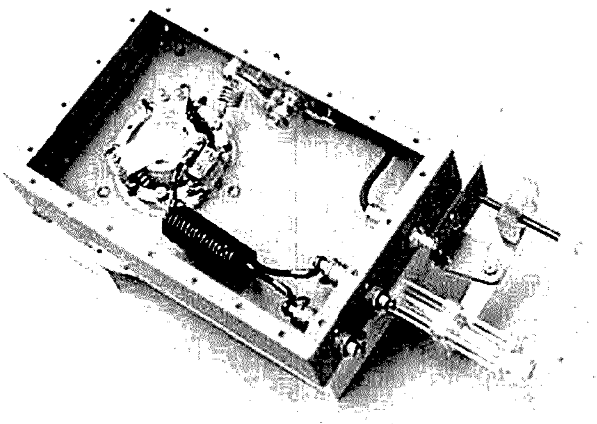

The complete amplifier is a look-alike to its high-frequency cousin, the Henry 2K.

## output loading circuit

The output loading circuit consists of a sliding tap on the plate line coupled to a 50-ohm, five-wire transmission line. The four outer conductor rods are fixed. The inner conductor is telescoping brass tubing, which allows loading adjustment by varying the position of the power take-off point along the plate tuned circuit. The inner conductor is 0.375-inch O. D. brass tubing, which is soldered at one end to the output coaxial receptacle. A second piece of brass tubing, 0.382-inch I. D., is telescoped over the fixed piece. To make good electrical

Al Roach, W6JUK, of Dymond Electronics, 515 Blackstone, Fresno, California 93701, is presently planning to market a 144-MHz rf deck very similar to the one in this amplifier, including the strip line, 3CX1000A7 tube and SK-870 socket. If there is sufficient interest he may offer a complete amplifier, including power supply, blower, metering and cabinet. If you're interested, write directly to him.

contact, finger stock backed up by a coiled spring is soldered to the larger conductor. The other end of the 0.381-inch I. D. tubing is inserted into a Teflon block, which slides along the four outer conductor rods. Finger stock is mounted on the Teflon block and



**Input circuits showing T-network and bi-filar-wound filament choke.**

soldered to the five-wire transmission line center conductor. The finger stock makes contact with the inside surface of the top slab of the plate tuned circuit, providing variable contact for adjusting the load. The center conductor rides in a Teflon sleeve bearing where the transmission line passes through the plate resonator

**Completed amplifier. Blower and filament transformer on reverse side protude into large empty space in top of cabinet.**

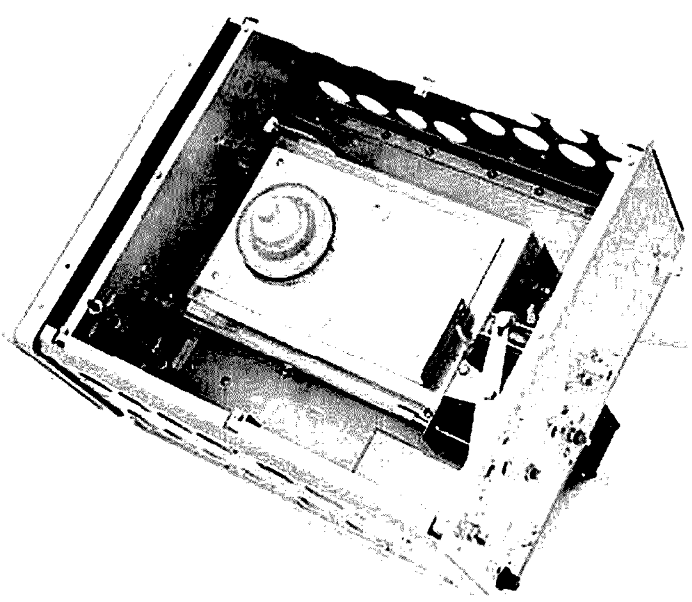


**table 1. Operating conditions for the 2-kilowatt amplifier.**

| | cw | ssb |
|---|---|---|
| plate voltage | 2900 volts | 2710 volts |
| plate current (zero signal) | 32 mA | 32 mA |
| plate current (single tone) | 332 mA | 690 mA |
| amplifier plus driver input power | 1000 watts | 2000 watts |
| grid voltage | -18 volts | -18 volts |
| drive power | 25 watts | 82 watts |
| filament voltage | 5 volts | 5 volts |
| filament current | 33 amps | 33 amps |
| power output | 500 watts | 1120 watts |

shorting block. The four outer conductor rods are threaded and screwed into tapped holes on each side of the shorting block.

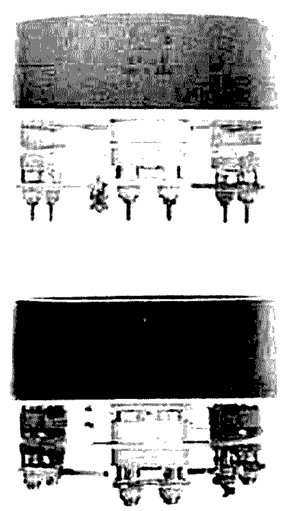
## performance

The amplifier operates over the 144-148 MHz band under the conditions shown in table 1. No intermodulation measurements were made, but previous experience indicates that third-order products will be 32 dB below one tone of a two-equal-tone signal.

### references


[1] Robert I. Sutherland, W6UOV, "Design Data for a Two-Kilowatt VHF Linear," *ham radio*, March, 1969, p. 6.


**The Eimac SK-870 socket before modification, below. Washers were removed to increase tubes self-neutralizing frequency, upper.**



# an electronic thermometer

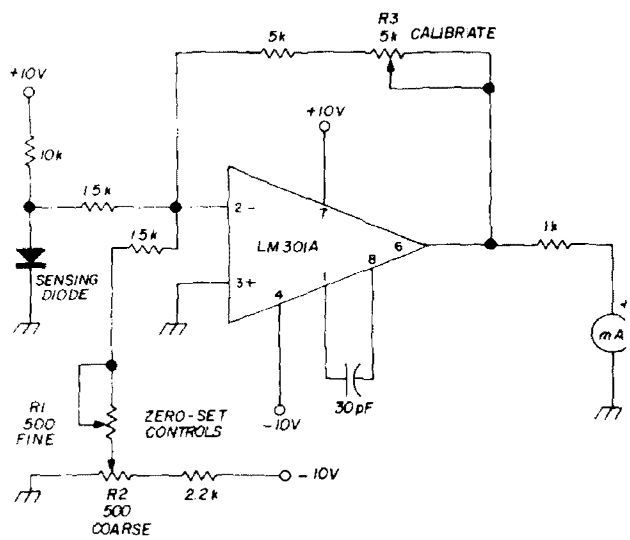
**A simple but effective instrument that can be built in just a few hours**

James Goding, VK3ZNV

**A forward-biased silicon diode** has a virtually constant negative temperature coefficient. The voltage across it drops about two millivolts per degree Centigrade (depending on the diode) over a wide temperature range.

This property has been known for years but has been seldom used for temperature measurement because of the complexities of the dc amplifier required for low drift and calibration stability. The

**fig. 1. Circuit for the electronic temperature indicator. The LM301A is a very stable dc amplifier.**



availability of inexpensive IC operational amplifiers has changed the picture considerably.[1]

The National Semiconductor LM301A operational amplifier, used in this circuit, costs less than six dollars. Its gain depends almost entirely on the feedback and input resistors, and the circuit is very stable.

## design

The electronic thermometer circuit is shown in fig. 1. The difference between the input currents to the op amp (i. e., currents through the two 1.5k resistors) is amplified and presented to the meter. This current difference is set to zero at 0° C by R1, R2; thus the meter indication is proportional to the temperature of the sensing diode. The calibration control, R3, sets the amplifier gain to suit the temperature coefficient of the diode.

## power supply

The instrument requires regulated +10 and -10 volts. Current drain is only about two milliamperes. A simple method of obtaining stable voltages is shown in fig. 2A. A couple of inexpensive 9-volt transistor radio batteries, connected as shown in fig. 2B, may be used for initial checkout of the thermometer. However, a regulated supply should be used for calibration and operation. The regulation of a battery supply isn't too dependable, especially after prolonged use when battery internal resistance increases.

## calibration

The zero point must be calibrated first. Place the sensing diode in melting ice, then adjust the zero-set pots for zero meter indication. Place the diode in boiling water, and adjust calibration pot R3 for full-scale meter reading. Bubbles of steam may cause the meter to fluctuate, but this can be remedied by allowing the water to cool until it just stops bubbling. The water temperature will then be about 99°C.

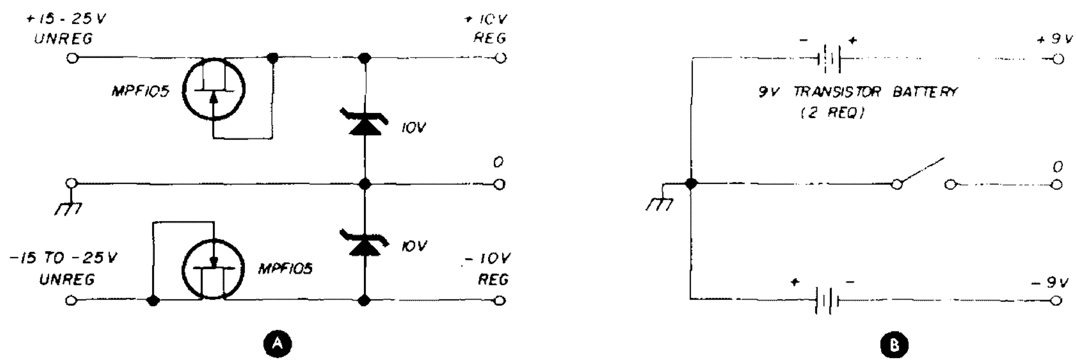



**fig. 2. Regulated power supply for the thermometer, A. Circuit in B may be used as an alternate for preliminary checks.**

## installation

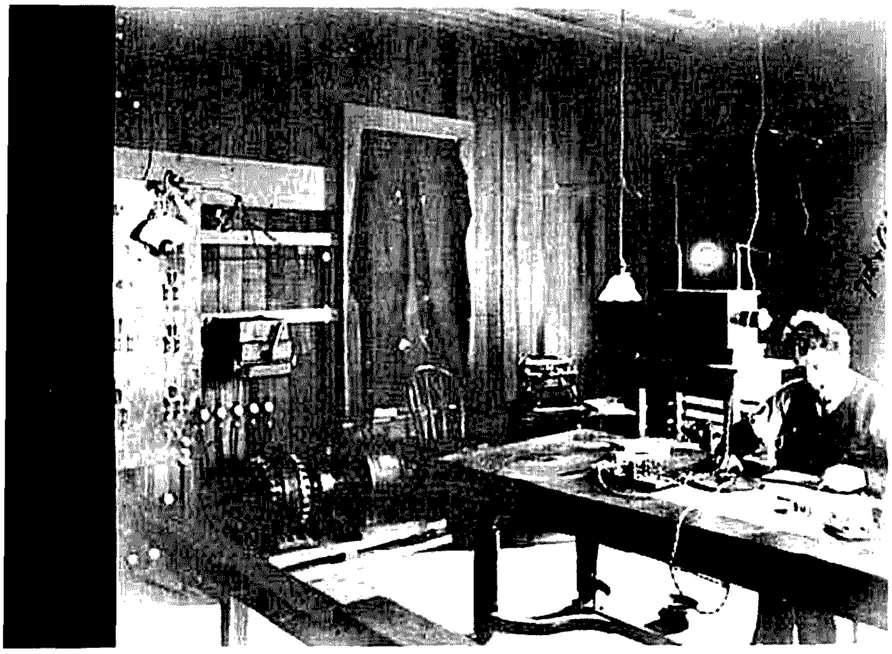
The instrument is now ready for use. The sensing diode may be placed remotely from the instrument. Long leads to the diode will have little effect on calibration. Thus the instrument is useful for measuring temperature in inaccessible places.

A suggested application would be a monitor for the TV sweep tubes used in many modern transceivers. These tubes can get pretty hot during prolonged key-down conditions in the cw mode.

### reference


1. Don Nelson, WB2EGZ, "What's This We Hear About Op Amps?", *ham radio*, November, 1969, pp. 6-23.


**ham radio**



Inside the PW station in Los Angeles, August 1903; Mr. Krenke on watch. The knife switches and "bug eye" meter were typical of a powerful wireless plant.

# catalina wireless 1902

Another glimpse into the early days of radio by an old timer who knew the old timers...

Ed Marriner, W6BLZ

Researching early wireless history is a part of ham radio I enjoy very much. It's a real challenge, because many events pertaining to new installations were not recorded in print. Newspapers didn't give much coverage to new wireless plants, probably because they were considered just another business venture without much news value.

Tracing the history of wireless communication calls for detective skills and infinite patience. To find really early information, one must talk to the oldest operator one can find, who might have talked to the oldest operator in his time. This takes a lot of digging and leg work, but results are rewarding if one has a soft spot in his heart for the early wireless pioneers.

Anyone who operated one of the very early stations is now nearing the century mark. He would be called an OOOT, or

Olde, olde, old timer. In descending order of venerability, early operators might be classified thus:

| | |
|---|---|
| OOOT | 1900 - 1910 |
| OOT | 1910 - 1920 |
| OT | 1920 - 1940* |

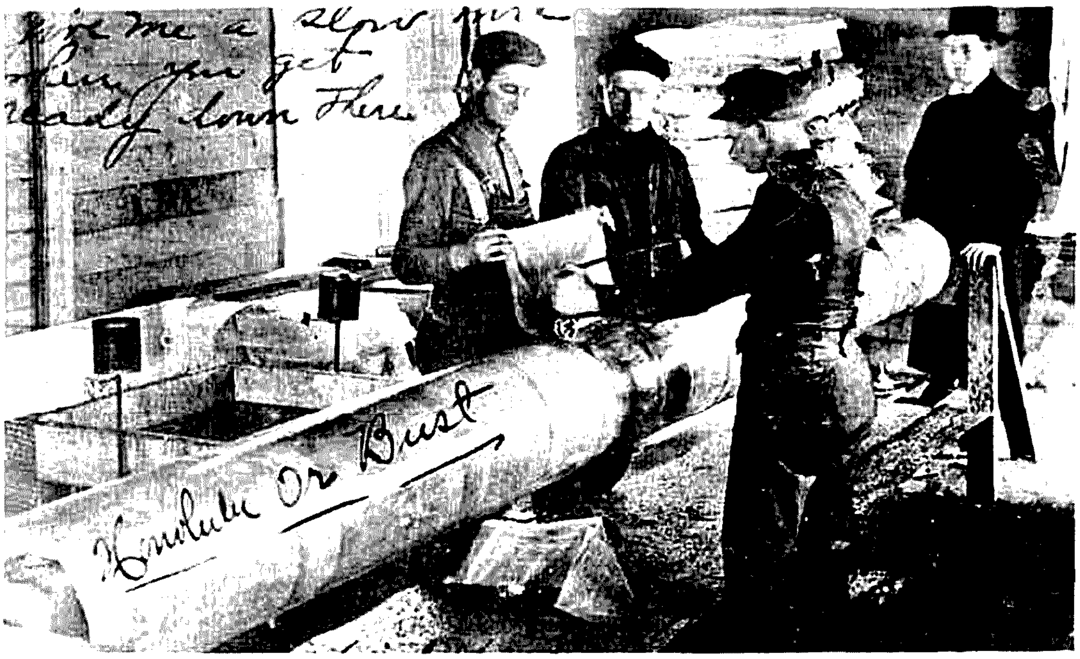
## detective work

One of the earliest wireless circuits I'd heard about was established by the Pacific Wireless Telegraph Company around the turn of the century. It was between Los Angeles and the city of Avalon on Catalina Island, which is about 30 miles off the California coast.

My quest for historical data on this installation took me through the Los Angeles Times' morgue, the public library, and the Avalon Chamber of Commerce. As I said, nothing of significance was in print so I kept looking.

The trail finally led to Howard Seefred, 6EA. Historians will recall 6EA's contributions to the early ARRL relay work in cross-country communications on the "short" waves. Howard has since passed away, but he provided me with some good clues about the old PW station and some of the operators. As a boy, Howard used to hang around the PW station at 7th and Alameda in Los Angeles. He recalled a PW operator by the name of A. F. Krenke, who gave generously of his time and knowledge to help 6EA get started in wireless.

Howard also remembered that Mr. Krenke had moved to San Diego, where he operated a station for PW until about 1916. With this information, I contacted old-time Navy operators and early hams in the area in an attempt to find Mr. Krenke. As if by a miracle he was located, and with a steady hand he described his part in the drama of early wireless in Southern California. This is his story.




PW technicians winding a helix coil for the Mt. Tamalpais station; Mr. Krenke at far left. Cloth stacked in corner was applied over each layer of wire and coated with shellac. Process was not unlike that used for making today's surfboards with fiberglas.

## early circuits

The Marconi circuit between Catalina Island and Los Angeles was installed in 1902. The power of the press was augmented by that of the 5-kW straight gaps of the PW stations. News items transmitted over this circuit were published in

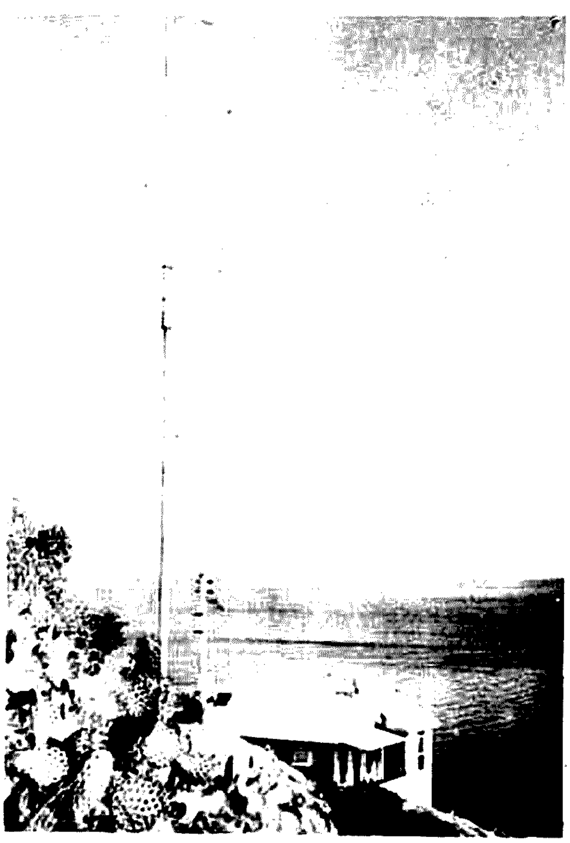
*The nonlinearity of the time scale after 1920 reflects the growth of the state of the art. Editor.

the Avalon paper. The Jefferies-Fitzsimmons fight results, for example, appeared in print hours before the news arrived by steamer from the mainland.

During the first few years of operation, the mainland PW station was located at San Pedro (15 miles or 15 minutes via the freeway today from downtown Los Angeles). In February 1905, the San Pedro plant was moved to 7th and Alameda after it was determined that messages could be sent the extra distance.

## more detective work

The first criminal case to be solved with the aid of wireless was recorded during the first year of operation of the PW circuit. Seems that two fellows left Avalon on the steamer with the cash and some liquid goods from the Hotel Metropole bar. The hotel manager, anxious to try the new wireless (and retrieve the loot), sent a message to the San Pedro police department.



Catalina Wireless, 1903. It's still a DXer's dream location. Mast was 70 feet high and supported a Marconi antenna.

The San Pedro police were waiting at the dock when the steamer arrived from Avalon. The culprits were apprehended, much to their amazement and chagrin. The cash was returned to the hotel manager, but the bottled goods somehow disappeared en route. In any event, this seems to be the first recorded evidence of public service by wireless telegraphy.

## help wanted

During this period, Mr. Krenke was a telegraph operator for the Southern Pacific Railroad at the Lone Palm Tree Watering Station (now known as Palm Springs).

Mr. Krenke went to Los Angeles shortly after the robbery and was fascinated by the excitement of wireless' contribution to law enforcement. Such events helped to increase business for PW, and they were looking for another operator. Mr. Krenke applied for the job and was assigned to the Avalon station, where he worked for about two years.

## catalina wireless

The Avalon station was on a hill above the dance casino (now the St. Catharine Hotel). Operating technique wasn't as polished as it is today. At the end of a transmission, for example, the operator had to shut down the motor generator so he could copy the message from Los Angeles. If he missed a word or two (which was not uncommon), he had to start up the gasoline engine that ran the generator so he could ask the Los Angeles station for "fills", as they were called. Such was break-in operation in those days. It was slow — but who hurried in 1903?

## the wireless news

A synopsis of the Los Angeles Times was transmitted to Avalon each day, amounting to about 600 words of press. The news, hot off the crudest of receivers, was typeset and printed in Avalon. The paper was called the "Wireless News" and sold for ten cents. A mint copy of this publication would be worth a great deal today.

Mr. Krenke, chief operator, Catalina Wireless, 1903. Blackboard at left announced incoming messages for the locals.

## business expansion

The Pacific Wireless Telegraph Company had big plans. Management made the decision to install a powerful station on Mt. Tamalpais, near San Francisco. The idea was to send messages all the way to Honolulu, an ambitious dream in 1906.

A construction crew from PW first installed a station in the Merchant's Exchange building, with offices on the 14th floor, in downtown San Francisco. Then a station and antenna tower were installed on Mt. Tamalpais.

On April 18th, 1906, an earthquake rocked San Francisco. The tower on Mt. Tamalpais came down, and the station was completely wrecked. This dashed the hopes of the company that tried so hard – and the date of wireless communication between California and Hawaii was set back many years.

## wireless san diego

Out of the San Francisco wreckage of 1906 emerged a new wireless company. It was called United Wireless and Telegraph, and was founded in 1908. Mr. Krenke worked for UW at their station in San Diego. The UW plant was installed on the Granger Building at 5th and Broadway, San Diego. This venture didn't last long, and Mr. Krenke again found himself out of a job. He delivered groceries for awhile to make ends meet, then he heard rumors of a new wireless station to be installed in San Diego. But I'll let Mr. Krenke finish the story.:

"I'd heard that the Federal Wireless and Telegraph Company was interested in establishing a station in downtown San Diego. I considered the possibility of the U. S. Grant Hotel as a site. The hotel manager, a Mr. Holmes, was sympathetic to the idea, and we wrote to the Federal Wireless Company in San Francisco.

"To make a long story short, I was hired by FWT's chief engineer and was put in charge of the station. I installed the antenna system. The transmitter was a Poulsen arc, which was powered by a 500-watt dc generator. I hired a messenger boy and was in business.

"After about five years (I'm not sure of the date, but I think it was 1915) a terrific storm hit San Diego. Heavy rain caused a dam to break, which created a flood that destroyed thousands of dollars worth of property.* San Diego was isolated except for the wireless station. During this terrible storm, my station handled Western Union and Postal Telegraph traffic in addition to our regular work. I stayed at the key for three days.

"There's little left to tell. I was sworn into the U. S. navy in 1917 and operated radio NPL on Point Loma. After World War I, I worked as a civilian operator in San Francisco, ending a half century as a wireless operator."

ham radio

*This was supposedly the result of work by one Charles Hatfield, an itinerant rainmaker, who was hired by the San Diego City Council to end a long drought. Hatfield never collected his $10,000 fee, because the flood was considered an "act of God". Litigation by Hatfield's descendants is still pending in San Diego courts.

# variable bandpass audio filter

One solution to the receiver selectivity problem is this RC feedback system featuring variable bandwidth to less than 50 Hz

Gary B. Jordan, 629 Manhattan Avenue, Hermosa Beach, California 90254

Many articles have been written to provide a solution to the selectivity problem existing in some popular amateur transceivers and older receivers. Most of these designs use a passive LC filter employing surplus toroids in a resonant circuit. The filter is usually an add-on device connected in the audio circuit.

While this is an acceptable approach, it does have some disadvantages. The resonant circuit at audio frequencies is bulky, and it's difficult to obtain high Q for good selectivity without some ringing.

Selective filters using RC feedback to obtain high Q without ringing haven't appeared often in the amateur literature. This article presents an effective solution to the selectivity problem using RC feedback techniques and provides some theory on basic principles. If you'd rather skip the theory, you can build the circuit shown in **fig. 6**, plug it in, and enjoy excellent variable bandwidth in a filter centered at 500 Hz.

## RC feedback networks

Passive elements in an appropriate circuit will yield voltage gain. Consider **fig. 1**, in which three resistors and three capacitors are connected in series-parallel. If C1 = C2 = C3 and R1 = R2 = R3, the network will resonate at a frequency Fo = 65/CR, where C is in microfarads and R is in kilohms. Applying an ac signal to terminals 1 and 2, a signal loss will occur at terminals 3 and 4. But at Fo, the signal at terminals 3 and 4 will be 180 degrees out of phase with the input signal. This is very important, as discussed later.

## three-terminal analysis

If the signal at the output were 10 times less than the input signal, the network would have a loss factor of 10. Another way of saying this would be that the beta (as in transistors, for example) is

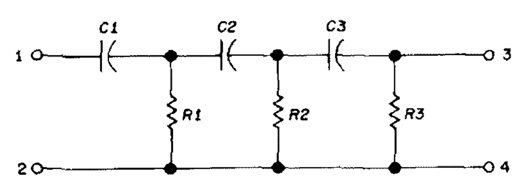
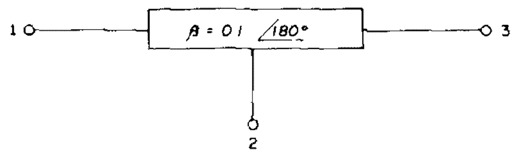



fig. 1. RC 180-degree phase-shift network.

equal to 0.1 at 180-degrees phase shift. The network could then be represented as in **fig. 2**, which is simply a three-terminal black box. If, as shown in **fig. 3**, a 10-V p-p signal were applied to the input, the output would be 1 V p-p precisely

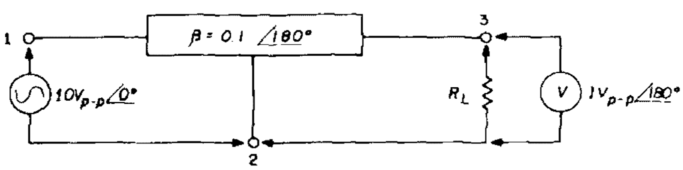
**fig. 2. Three-terminal equivalent of the RC network.**




180-degrees out of phase with the input. When the input wave reaches a maximum, output would be minimum, and vice versa.

Now consider something quite interesting. If a voltmeter were connected *across* terminals 1 and 3 and we consider only one instant of time in the ac waveform of **fig. 3**, that instant will be

**fig. 3. RC network input/output relationship.**




the precise point at which the input reaches a maximum of 10 volts and the output reaches a minimum of -1 volt, since the output is 180 degrees out of phase with the input. Looking at **fig. 4**, we see that input and output ac signals have been replaced by two batteries representing the selected instant of time. Hence, a 10-volt battery represents the peak input, and a 1-volt battery the phase-reversed output.

## voltage gain

The really interesting feature is that a voltmeter connected across terminals 1-3 will indicate 11 volts. This is a higher voltage than either input or output because it is the *sum* of these voltages. For every point on the ac wave, the voltage across terminals 1-3 will always be greater than that across terminals 1-2. The reason is that the phase shift causes the output at terminals 1-3 to be the sum of voltages from terminals 1-2 and 3-2. A voltage gain therefore exists at terminals 1-3 with respect to terminals 1-2.

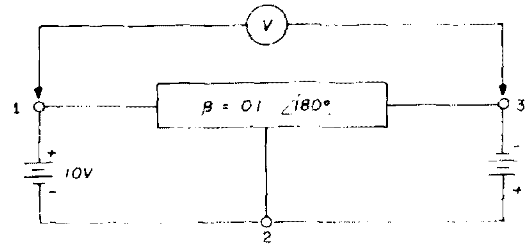
To take advantage of this gain, we select 1-2 as the input as before, but take our new output from 1-3, as shown in **fig. 5**. Thus, we have created an ac transformer of sorts that has a voltage step-up, but at only one frequency: that at which 180 degree phase shift occurs.

## RC audio filter

A selective audio filter may be created by connecting the phase-shift network around an amplifier so that the network's voltage gain is utilized. If the product of the amplifier gain, K, and the new network's beta, B, is greater than unity, the circuit will oscillate. However, if K is made less than unity, the circuit won't oscillate but will have an extremely high Q at the feedback network's resonant frequency. The system's bandpass will become narrower as the product of B and K approaches unity.

To implement this idea, it's necessary to find an amplifier with less than unity *gain. An emitter-follower fills the bill* quite well, since its K is 0.98 typically. **Fig. 6** shows the RC network connected in the voltage-gain mode around the emitter-follower. R6 controls feedback to a point below where oscillation would

**fig. 4. View of the network at one instant of time.**




occur. Thus, R6 provides a means of controlling bandwidth.

You might wonder what happened to the third resistor in the phase-shift network. It is still there, but as a virtual resistor composed of a parallel combina-

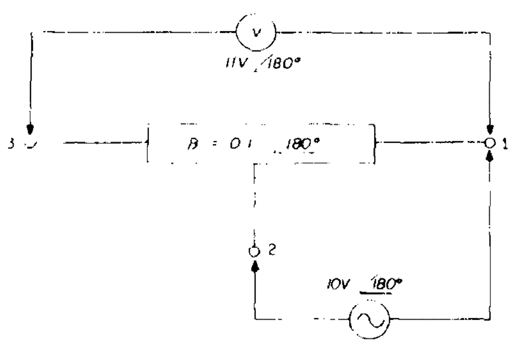


fig. 5. Reconnecting the input and output to yield voltage gain.

tion of all the resistance at the transistor base, and B times all the resistance at the emitter. In network notation, the virtual resistor consists of

$$R3 \| R7 \| B \quad (R10 \| R8 + R9)$$

The bias is obtained through R7, which provides dc bias as well as a small amount of degenerative feedback for stability.

tion is needed because the voltage at C1 must be kept to a few hundred millivolts p-p to avoid over-loading the circuit.

Higher inputs will cause some loss of selectivity. Resistors R1 and R2 may be selected to reduce the audio from, say, a receiver's headphone output to the proper level. Typical values of R1 and R2 would be about 2K and 200 ohms respectively.

## construction and operation

Construction isn't critical, but care should be taken to separate input and output circuits, and the supply voltage should be well bypassed at audio frequencies. Typical supply voltage would be 6-15 volts dc; a 9-volt transistor battery could be used.

Varying R6 will allow a passband between several hundred Hz to less than 50 Hz for copying cw signals. If R6 is

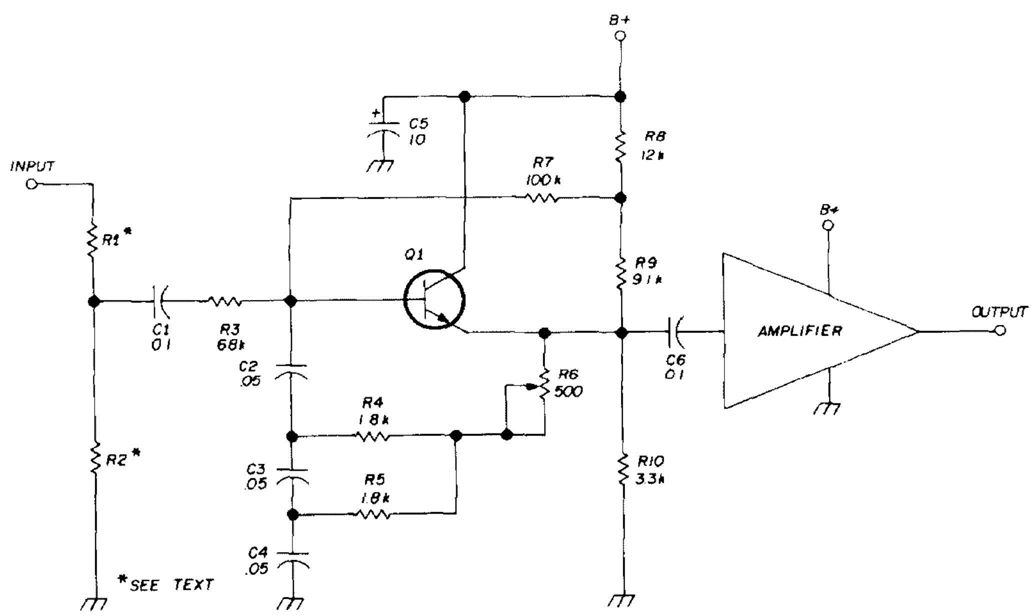


fig. 6. Variable bandwidth selective audio filter. Capacitor C5 is a tantalum electrolytic; other capacitors are disc ceramic. Transistor Q1 is any small-signal npn transistor. The amplifier can be any inexpensive audio amplifier IC.

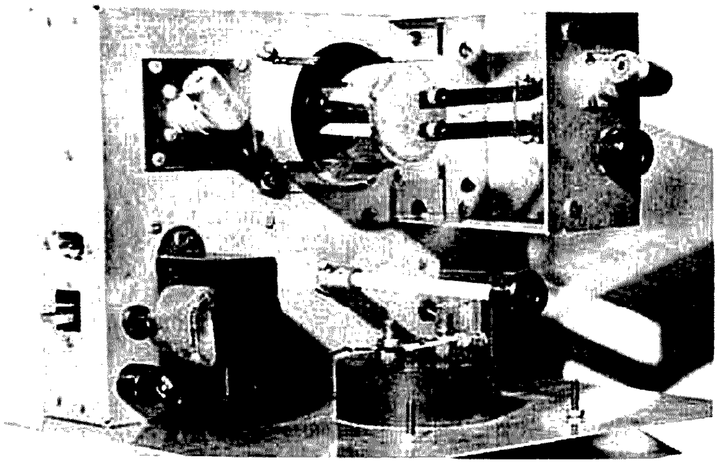
## the second amplifier

The amplifier shown as a block in **fig. 6** may be any of the inexpensive, commercially available audio amplifiers in potted form, a microcircuit, or a simple home-built audio stage. Some amplifica-

operated near the shorted-out point, the circuit will oscillate.

Considering the circuit's simplicity and the excellent results, this seems to be an easy way to obtain audio selectivity for true single-signal reception.

**ham radio**



# rf power amplifier for 432 MHz

A design featuring resonant-line tank circuits and class C or AB1 operation

James W. Brannin, K6JC, 424 Anson Avenue, Rohnert Park, California 94928

If you would like to boost your signal on 432 MHz (and who wouldn't?), this amplifier will do the job with only 4 watts of drive. It provides 100 watts input on cw and ssb and 65 watts on a-m. A variable bias control allows instant selection of class C or AB1 operation. A type 5894 tube is featured. The 5894 performs well in the 400-MHz band and is available at reasonable cost from dealers in used uhf equipment.

## the circuit

The tuned-circuit elements are conventional. A ¾-wavelength grid tank and series-fed push-pull ¼-wavelength plate tank are used. The circuit is shown in fig. 1. Bias can be varied from -60 to -28 volts for Class C or AB1 respectively. I've had no problems with parasitics or self-oscillation on either mode.

## construction

A 9 x 7 x 2-inch chassis accommodates all components, including transformers and bias supply. Construction details for grid and plate lines, as well as for the

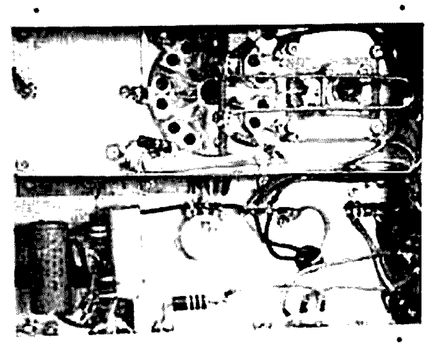


plate tuning capacitor and output link, are shown in **fig. 2.**

Shielded wire is used below chassis to minimize stray coupling. Coupling between input link and grid tank should be fairly close. The loops should be spaced about 3/16 inch.

The 5894 socket preferably should

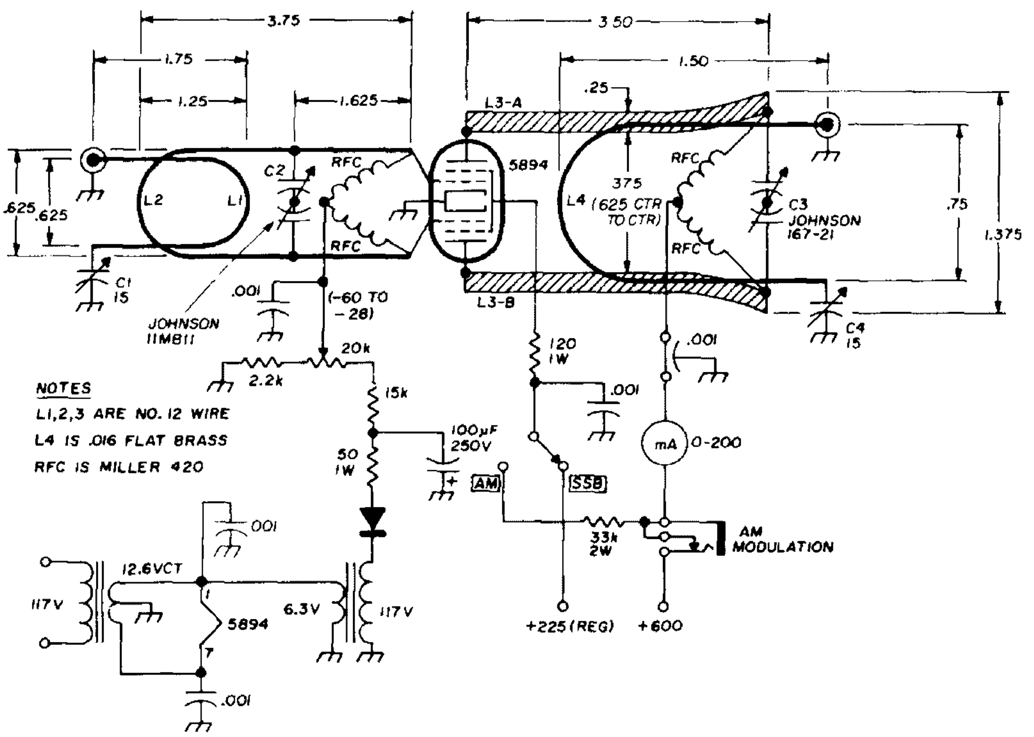



**fig. 1. Complete circuit details for the 432-MHz rf power amplifier. Screen and filament bypasses may be built into socket. Idle plate current 40 mA normal; 100 mA peaks on ssb.**

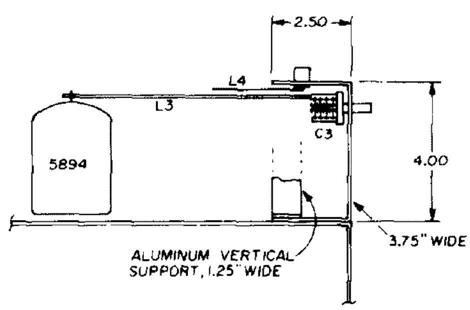
have built-in bypass capacitors. A conventional socket will work, however, if the leads to the socket bypasses are *short.*

## tuning and adjustment

You won't have any difficulty tuning the amplifier if inductance dimensions are followed. Some kind of output-power indicator is required. If you don't have an in-line swr/power meter, an acceptable indicating device can be made as follows.

Sew about two inches of enamelled wire beneath the output coax cable shield. Terminate one end of the wire in a noninductive 50-ohm load on the end toward the antenna. Connect the other end of the wire to a 1N34 diode. Bypass the diode, and connect the other end of the diode to a 0-200 microammeter. (If you use a 0-1 milliammeter, the pickup wire length will have to be increased to about six or eight inches.)

Decrease plate and screen voltage to about 300 and 150 volts, and tune the amplifier for maximum output into the 50-ohm load. Disconnect the dummy




load, connect your antenna, apply full power, and you're in business.

After initial adjustment, little or no retuning will be necessary unless a fairly large change in operating frequency is made.

My 5894 amplifier has been operating for several months and has put a signal into the fringe areas over 100 miles away across several mountain ranges.

**ham radio**

# improving overload response in the Collins 75A-4 receiver

Simple modifications provide 13 dB higher signal-handling capability in this fine receiver

Raymond F. Rinaudo, W6ZO

Having obtained only limited results trying to improve the performance of commercially designed electronic equipment, I came to the conclusion that the designers knew what they were doing all along. Even so, the temptation to modify equipment is hard to resist, and my latest modification attempt resulted in a 13-dB improvement in strong-signal-handling capability of the 75A-4 receiver. However, this was achieved by second-guessing Collins engineers some fifteen years after the set had been marketed!

A recent book on ssb techniques indicates that a substantial improvement in receiver overload response can be obtained by reworking the mixer circuits.[1] Those interested in the subject are urged to obtain a copy of this work.

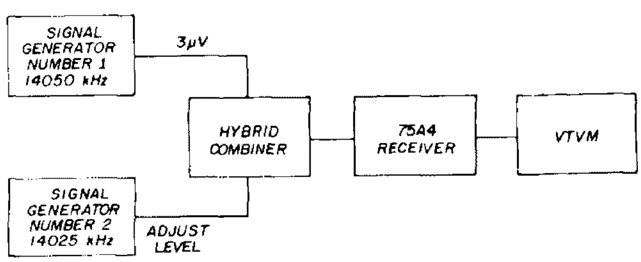
I decided to modify my 75A-4 because several nearby stations caused front-end overload. This causes a decrease in weak-signal strength, even though the interfering signals are 25 to 50 kHz away. It's all but impossible to copy a weak cw signal under these conditions.

The remedy is easy: about five dollars worth of parts and a little time. Here's how to do it.

## primary work

First replace marginal tubes, then align the receiver. Overload response may be checked using the setup in fig. 1. Set signal generator 1, representing the desired signal, to 3 microvolts at 14,050 $MH_Z$. Turn the receiver avc off and the bfo on. Now increase generator 2's level (at 14.025 MHz) until the desired signal, measured by the vtvm, decreases by 3 dB. This will require about 13,000 microvolts. All subsequent measurements are referenced to this level.

**fig. 1. Test circuit for measuring effectiveness of receiver modifications.**



## first mixer modifications

Replace the 6BA7 first mixer with a 12AT7. The modifications, fig. 2B, require only one 470-ohm ½-watt resistor and the 12AT7. Remove R14, R15; C35, C36. Revise the heater circuit as shown.

Next replace the 100-pF coupling capacitor with a 15-pF silver mica, then connect another 15-pF silver mica between grid and ground. This forms a capacitive voltage divider that reduces signal level to the first mixer grid. (The photos show a modified and unmodified set; actually, these are photos of two different receivers.)

After these changes have been made, peak the mixer grid and crystal-oscillator circuits for each band; also peak the mixer plate circuit. Only the capacitors should be peaked: C23, C26, C28, C30, C31, C32 and C17 in the mixer grid and C53 in the plate circuit.

The high-frequency crystal oscillator tuned circuits must also be retuned. Peak the tuning slugs on L11 through L17. Until the oscillator has been peaked, you'll probably find that the 21-MHz and higher-band crystals won't oscillate.

After the first mixer has been modified, a 24,000 microvolt signal will be required to cause a 3-dB decrease in the desired signal. I made this test before the two 15-pF capacitors were installed.

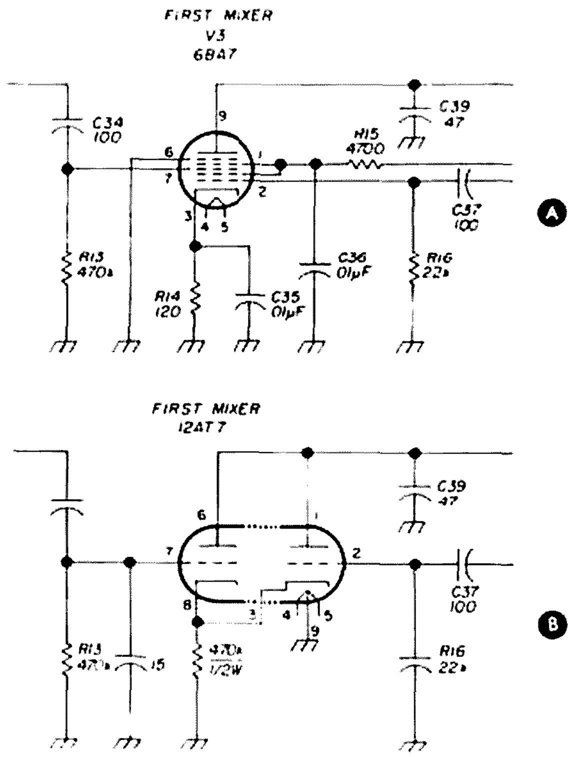
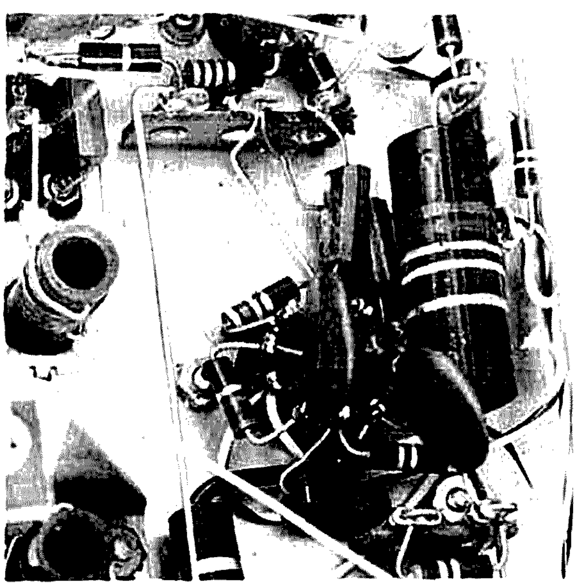



fig. 2. The 75A-4 first mixer before modification, A, and after, B. New parts consist of the 12AT7 and the 470K resistor.

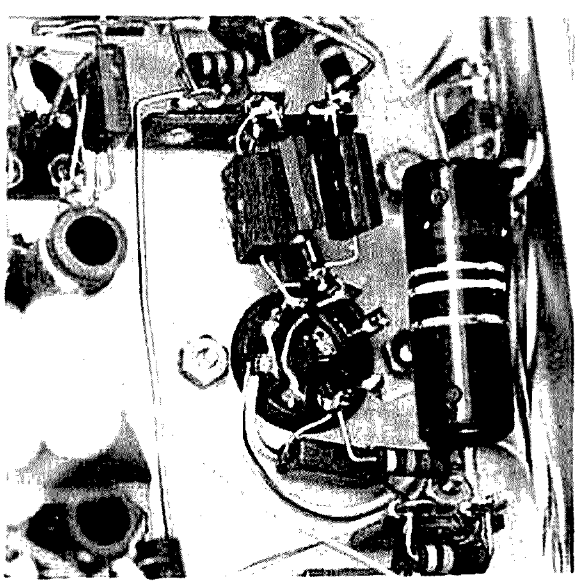
## second mixer modifications

The next step is to replace the 6BA7 with a 6DJ8. Before and after circuits are shown in fig. 3. The new tube plus four new parts are required: a 1k, 1.2k and 3.3k ½-watt resistor and an 820 pF silver mica capacitor. If you can't find an 820 pF capacitor, anything between 680 and

Wiring of unmodified 75A4 first mixer.



Modified first mixer; note reduced number of parts.



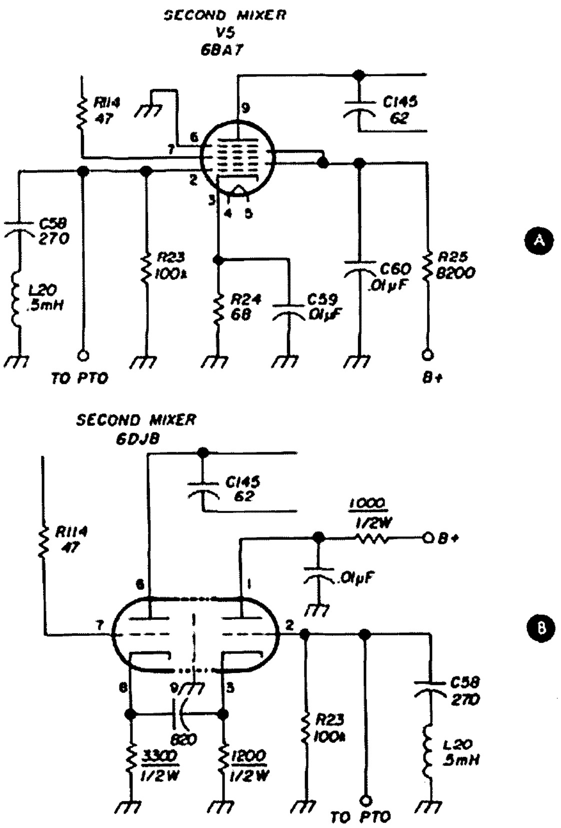



fig. 3. The 75A-4 second mixer. Original circuit is shown in A; modifications in B. Only five new parts are required for the change.

1000 pF will do. Remove R24, R25 and C59. Heater wiring is unchanged in this circuit. Complete the revision shown in **fig. 3B**. Peak the grid input circuits by tuning C56. Do not change inductor tuning.

After making the changes to the second mixer, check receiver performance again. A 40,000 microvolt signal from generator 2 will be required to reduce the desired signal by 3 dB.

## final checks

In my receiver, the capacitive voltage divider was installed after modifications to the second mixer. A final check showed that an undesired signal of 60,000 microvolts was required to reduce the desired signal by 3 dB (voltage ratio of 4.5).

## conclusions

Measurements at 28 MHz showed an improvement of 3 dB in signal-plus-noise ratio. Over-all receivers gain was somewhat lower, but this was more than compensated by the receiver's response to weak signals in the presence of local signals. More than enough gain was still available, however.

A type 6922 tube can be used instead of the 6DJ8. The 6922 has the same characteristics as the 6DJ8 but costs about three dollars more.

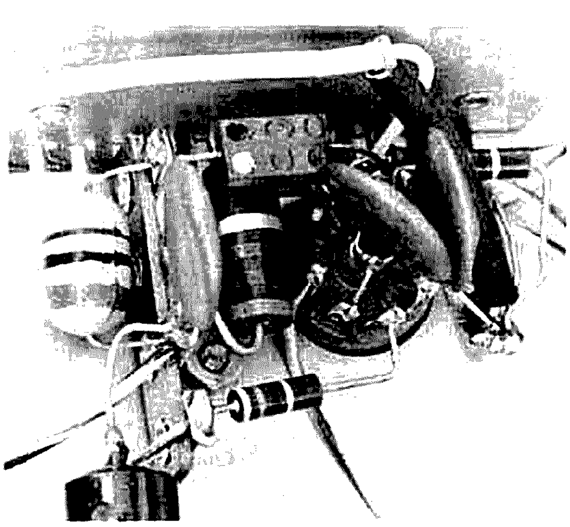
The modifications described are easily applied to the Collins 75A-4 receiver. The improvement in performance is well worth the investment in time and money.

references

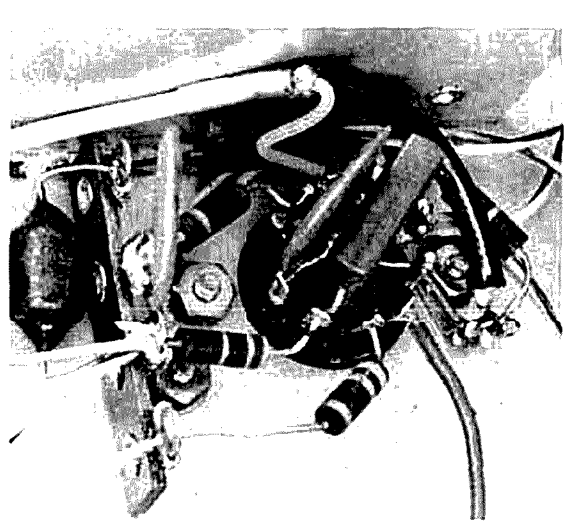
[1] Pappenfus, Breune, and Schoenike, "Single Sideband Principles and Circuits," McGraw-Hill, Inc., New York.

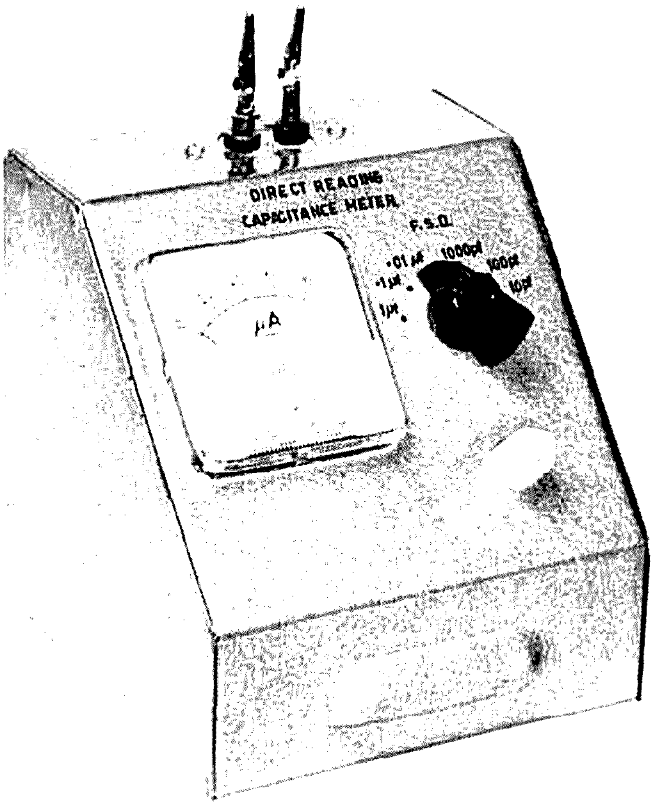
**ham radio**

Unmodified 75A4 second mixer.



Component layout of modified second mixer.






# direct-reading capacitance meter

An easily built instrument with many uses around your station

Mark King, ZL2AUE

The instrument described here evolved from a bench lashup I used for measuring small values of capacitance. The original circuit used a cross-coupled multivibrator to provide the required square wave. I developed the present circuit to avoid complicated switching and to minimize the number of parts.

The meter has six ranges. The lowest is 0-10 pF; the highest is 0-1 μF. The scale is linear, and if the timing capacitors are chosen carefully, the meter should have at least 5 percent accuracy. This is better than the tolerance on many capacitors between 1-100 pF.

## features

I've found this instrument to be invaluable for measuring:

1. Tuning capacitance–maximum and minimum values
2. Coax cable capacitance
3. Feed-through capacitance values (important at vhf)
4. Circuit strays
5. Capacitors with obliterated markings
6. Junction diode reverse capacitance

The instrument is also useful as a square-wave source for test purposes.

## operating principles

If a square wave is applied to a capacitor and resistor in series, and if their time constant is much shorter than the square wave's period, the combination acts as a differentiator, which produces a sharp voltage spike. If these spikes are integrated, a voltage will be developed that has a linear relationship to the square wave's period if the period is varied over a small range.

If, instead of varying the square wave's period, the differentiating capacitor's value is varied over a limited range, a similar effect occurs wherein the integrated voltage bears a linear relationship to the differentiating capacitance.

## practical considerations

In practice the unknown capacitor provides the differentiating capacitance, and the meter movement acts as an integrator. The meter also provides the resistance for the differentiator. This being the case, a 0-100 microameter was chosen, because its high resistance (in this case 10k) made for easy differentiation of small capacitances at a relatively low square-wave frequency. This should be born in mind if you wish to use a meter with a lower sensitivity.

On the other hand, the largest capacitance that can be measured with a low square-wave frequency is limited by needle jitter. The meter's integrating effect decreases as you approach dc.

## design

Initial mockups of the circuit used a unijunction transistor to generate a sawtooth wave that was applied to a limiting amplifier. The problem with this scheme was that, at the higher frequencies, insufficient output was available from the transistor at hand to drive the limiting amplifier.

In practice it proved less expensive to use the transistor pair in fig. 1, and it was also easier to adjust this combination for optimum waveform. It was pure coincidence that the timing capacitors corresponded to the maximum capacitances of the various ranges. The pre-set pots in series with the meter on each range are for very fine adjustment only. Too much resistance here will disturb the linearity of each range. If the calibration can't be brought within range with the pots, then the timing capacitors must be altered.

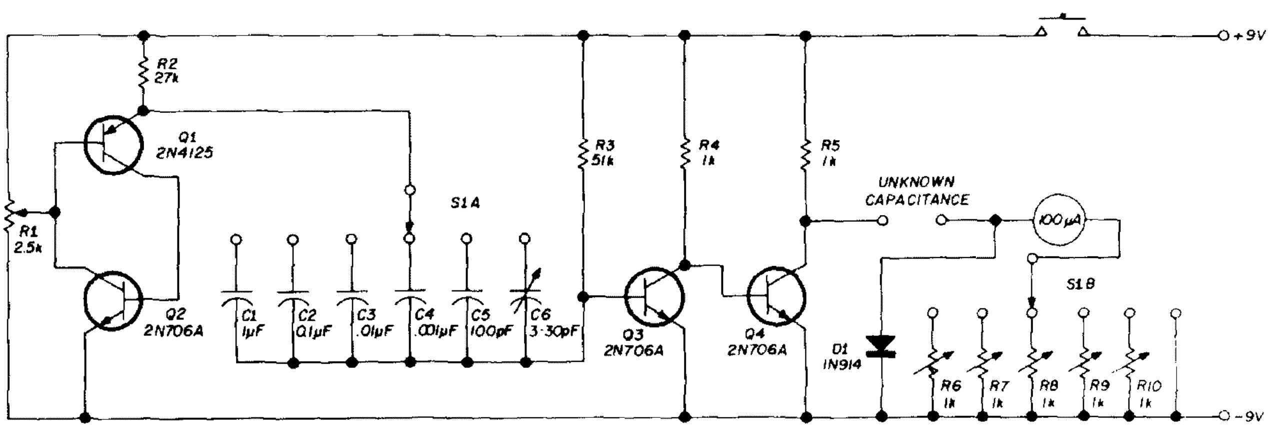



**fig. 1. Capacitance meter schematic. R1 adjusts output of square-wave generator Q1, Q2. Trim pots R6-R10 adjust timing capacitors C1-C6 during calibration, then remain fixed. Instrument should provide at least 5 percent accuracy.**

## construction

All components except the range switch, pushbutton switch, and meter are mounted on a PC board. The printed circuit was masked-in with a marking pen applied directly onto the laminate board. Unwanted copper was etched away with a concentrated solution of ferric chloride. The board was floated on the solution. Etching took about an hour at room temperature.

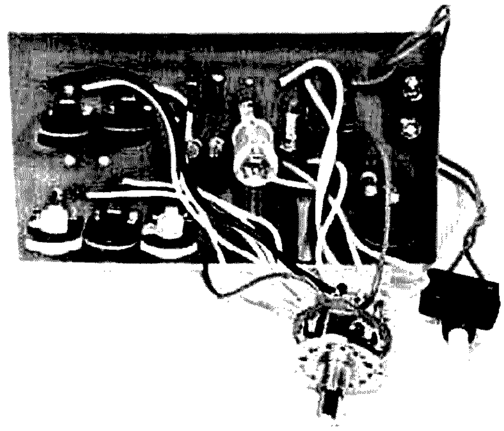
When completely etched, the board was thoroughly washed and dried. Remaining marking-pen ink was washed off with isopropyl alcohol; methyl alcohol will work as well.

The battery terminals were salvaged from an old type 216 transistor-radio battery. They were mounted directly on the circuit board with the other components. The measurement terminals were made by soldering small alligator clips to banana plugs. Sockets were mounted through a piece of lucite. They are spaced ½-inch apart. A hole was cut in the top of the instrument enclosure to clear the terminals by ¼ inch.

## calibration

Calibration accuracy is determined by the tolerance of the timing capacitors. If

"Ten bucks for the crystals -- and 50 cents for the rig."



Parts layout. Chassis is an easily made etched PC board, described in the text.

possible, choose values in the middle of each range, and adjust for correct readings with the trim pots.

If calibration can't be obtained with the trim pots, then the timing capacitors will have to be adjusted. As a further check, use a capacitor that gives a full reading on one range. Then the meter should read only 10 percent of the next-higher capacitance range.

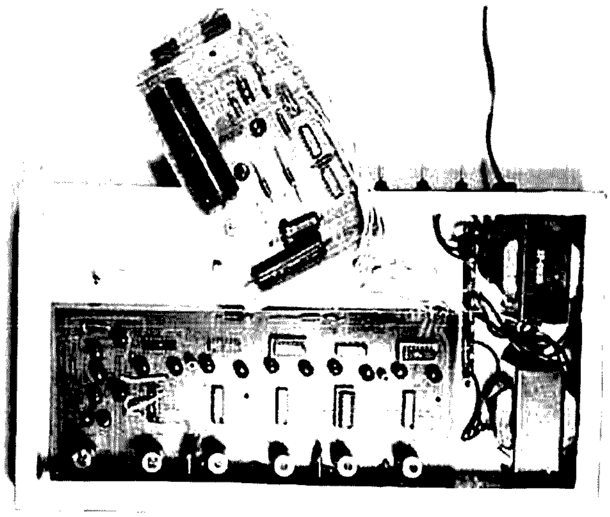
## operation

When measuring a capacitor, set the range selector to the highest range for a start. If the capacitor under test is larger than the range maximum, the capacitor will act as a low-impedance coupling between the drive transistor and the metering circuit, thus overdriving the meter.

The maximum current through the meter on the lowest capacitance range, when a large-value capacitor is connected, is 3 mA. If the terminals of the instrument are short-circuited, the current will be only 60 microamps. This is because Q4's collector will be effectively grounded through the metering diode. If the meter tends to read in a reverse direction, this indicates a leaky capacitor being tested.

The instrument is powered by a 9-volt transistor battery. Current drain is about 20 mA, but battery life should be good since a pushbutton switch connects the battery only when required.

ham radio



# 24-hour digital electronic clock

Osburne C. Stafford, K4ALS,

interesting in building an all-electronic digital 24-hour clock? With some of the new inexpensive digital integrated circuits that are on the market you can put one together for less than $230 if you buy all new parts. When Motorola Semiconductor announced their MC780P decade counter and MC9760 decoder a few months ago, it was like putting a good nickel cigar on the market. This clock is designed around these ICs, as is the optional built-in ten-minute timer shown in fig 2 Further cost savings resulted by using Burroughs B-5750 Nixie tubes—just about the most inexpensive readouts available.

Although a 24-hour clock is easier to build than a 12-hour version, all of the ingredients for a 12-hour version are included in the 24-hour clock with the exception of a couple of low-cost gates.

In the 24-hour clock, fig 1 , the "seconds" and "minutes" section count to 60, and then reset to zero. At the end of each hour a pulse is transmitted to the decade counter in the "hours" section. This continues until 23 hours, 59 minutes, and 59 seconds (23:59:59). The next pulse from the clock generator in the power supply resets the "seconds" and "minutes" counters to zero, and advances the "hours" section to 24. The number "24" is recognized by the decoder made up of gates G1 and G2, triggering the Schmitt-trigger circuit that resets the "hours" section to zero.

## clock pulse generator

The power supply and clock pulse generator are shown in fig 3 The one-second pulses are derived from the ac power line. The 60-Hz line signal is applied to integrated circuit U1 which is connected as a Schmitt-trigger: the output is a fast-rising pulse that is used to trigger the divide-by-six counter, U2; integrated circuit U2 drives a divide-by-ten stage, U3. Switches S1, S2 and S3 are used to select a divided output so the clock can be quickly set manually. Capacitor C1 suppresses spikes on the line and keeps rf energy from falsely triggering the solid-state circuitry. The output of the divide-by-ten counter is buffered by one half of U1.

A regulated supply of 3.9 volts is required for the ICs; the 170-volt output

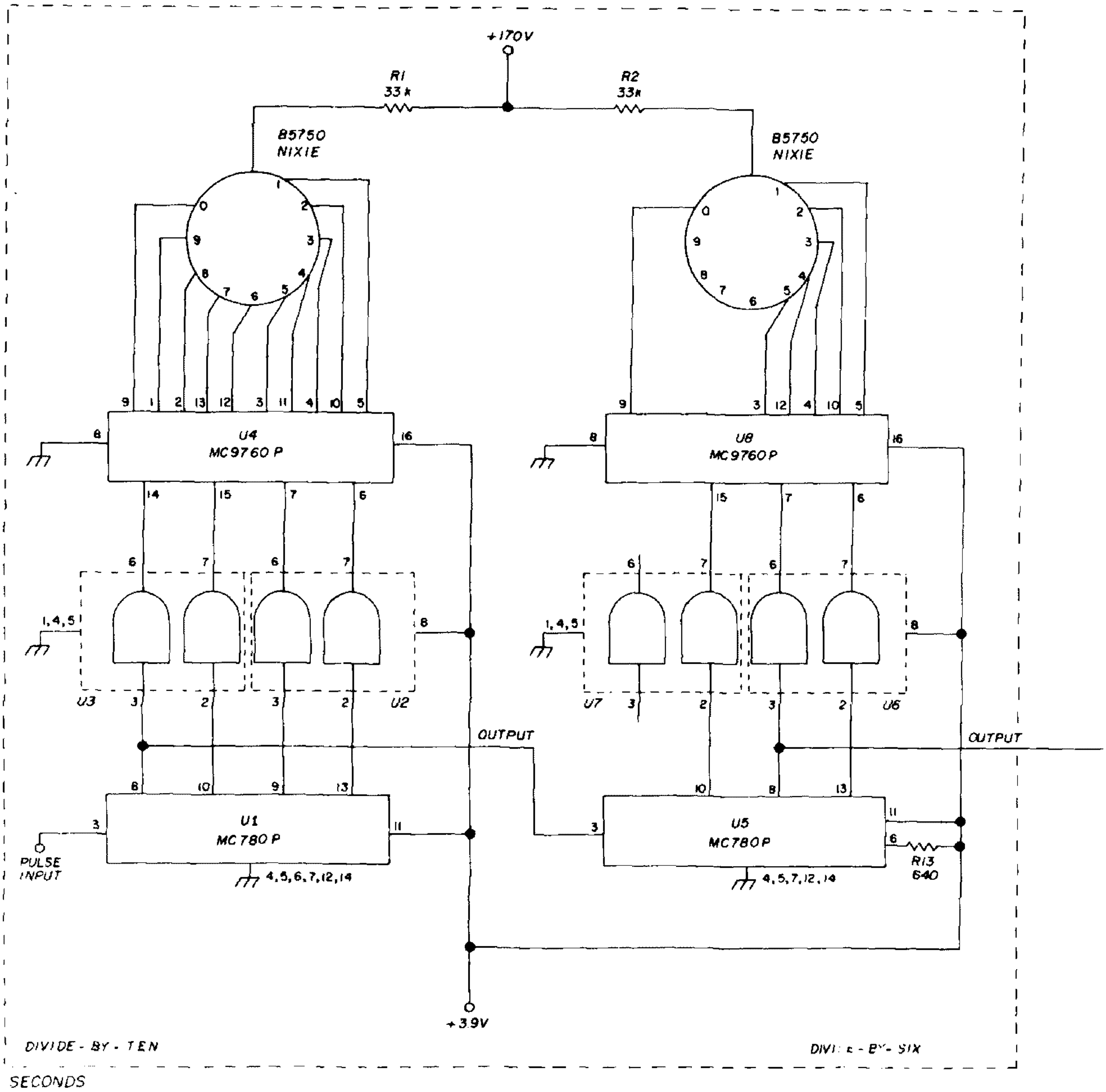



**fig. 1. Complete circuit diagram of the all-electronics digital 24-hour clock.**

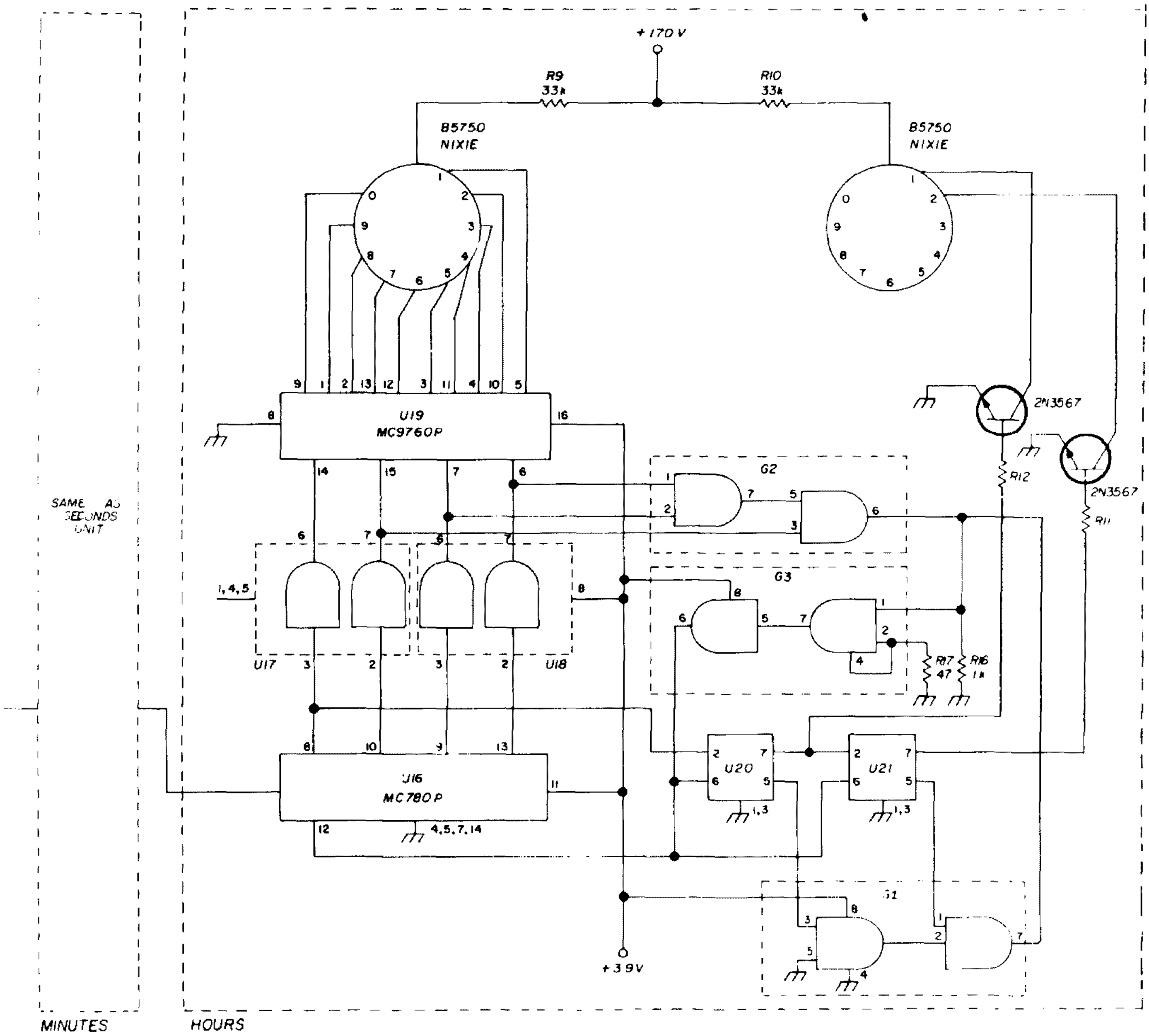
is for the readout tubes. This should be checked closely after the power supply is put together because resistors R1, R2, R5, R9, R10 and R18 were designed for 170-volt operation. The MC9760P will handle 70 volts, but it's best to start with the voltage on the low side (with dimmer readouts) than to zap the expensive integrated circuits.

## construction

Etched circuit boards are used throughout the clock shown in the photographs.* The completed circuit boards are mounted in a 13x7x2-inch aluminum chassis. On one side a rectangular cutout, 1½ by 8¼ inches, allows the readout tubes to be seen. A thin sheet of tinted plastic covers the front of the clock. The switches for setting it are mounted on the rear deck next to the ac line cord.

The colons between hours, minutes and seconds are formed by miniature neon lamps, their bodies painted except for the tips. The wire leads to the lamps are left about two inches long, and formed to hold the lamp in the correct position.

*circuit boards for this clock are avaiable from Stafford Electronics, Inc., 427 So. Benbow Road, Greensboro, North Carolina 27401. Clock board, DCU-700, $12.50; power supply board, DCU-720 $2.75; 10-minute timer board, DCU-721, $3.00. A complete kit of electronic parts for a 24 hour clock (less 10-minute timer) is $230.00.




## circuit variations

If you want 12-hour operation instead of 24, make the following circuit changes:

1. Remove R11, Q2 and U20.
2. Cut leads 2 and 7 on G1 and put a strap between these two points.
3. Remove the strap from pin 12 of

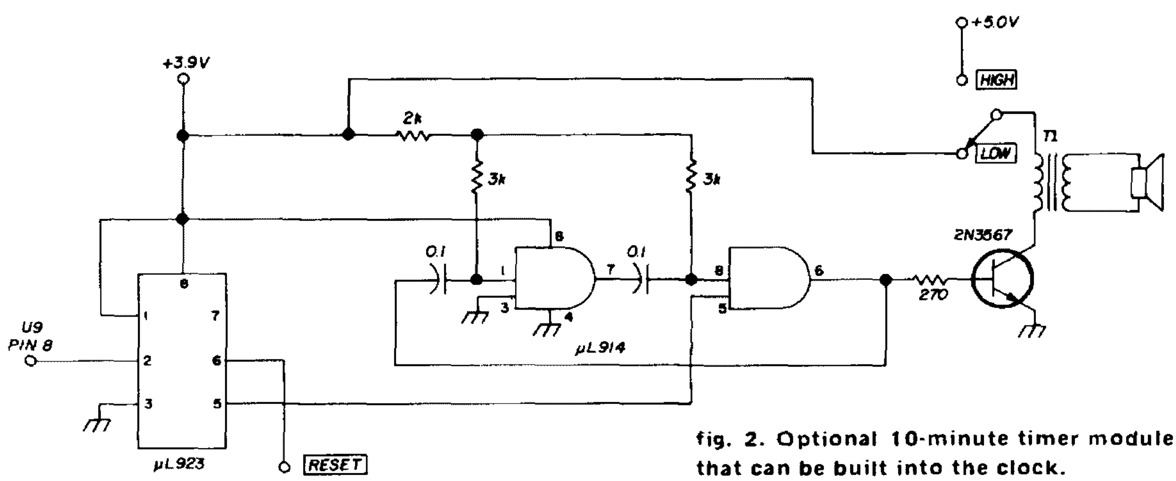



fig. 2. Optional 10-minute timer module that can be built into the clock.

U16 and the reset line, and the strap between pine 14 and ground. Connect a strap between pin 12 of U16 to pin 1 of G3 and connect pin 14 of U16 to the reset line.

Hz power, this clock can still be used. Simply convert the divide-by-six counter in the power supply to a divide-by-five stage—tie pin 14 of U2 to pin 3 and remove R8.

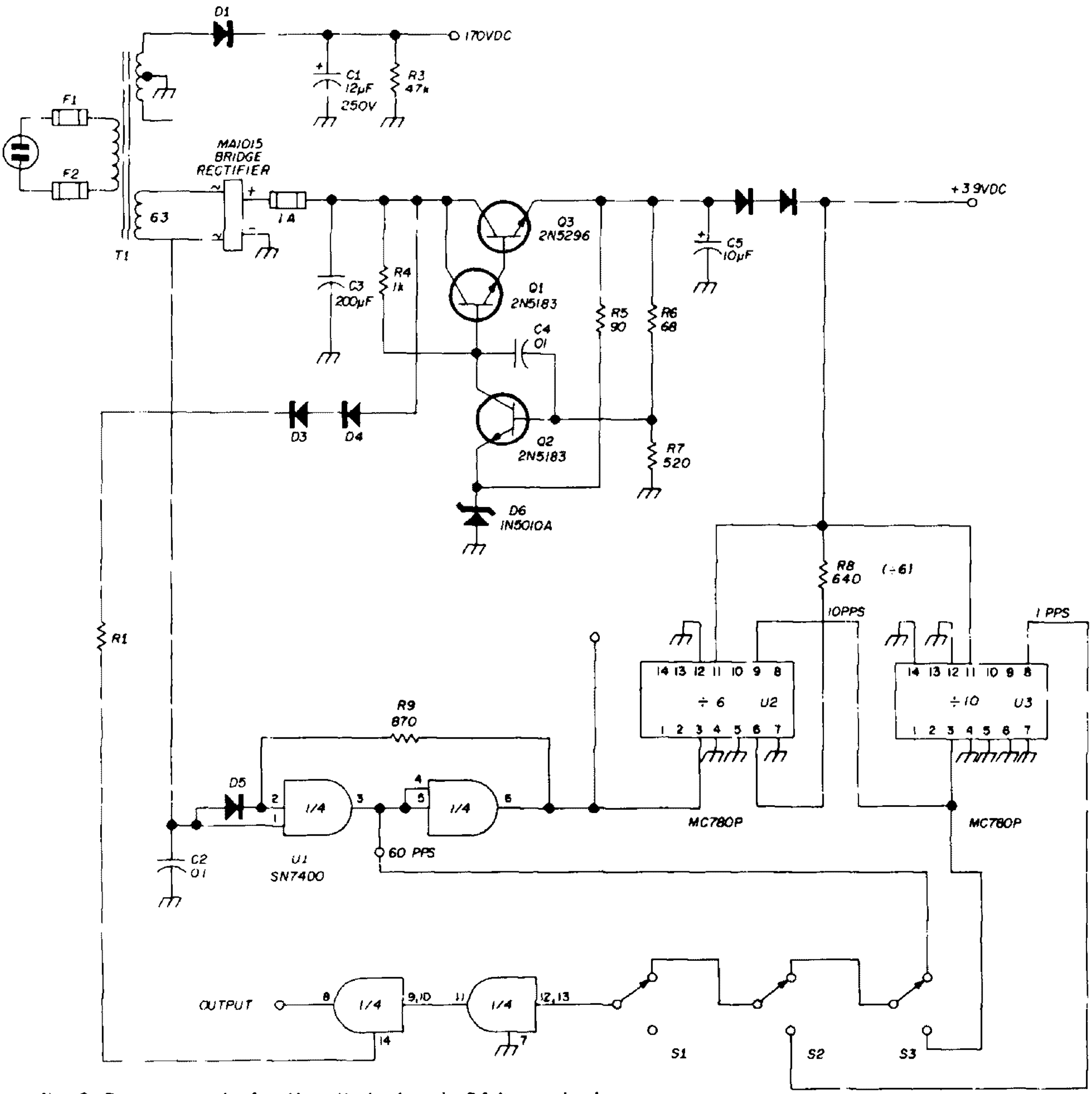



fig. 3. Power supply for the all-electronic 24-hour clock.

4. Add a 0.1 μF capacitor from pin 7 of G3 to ground.

With these changes, when pins 6 and 7 of U11 are at zero (indicating one), then pins 1 and 6 of G3 go high, resetting U16 to "1" and U19 to zero; the clock now reads 1:00:00.

If you're overseas, and serviced by 50-

After seeing a clock of this type in operation, you're practically compelled to build one. It's guaranteed to stir up some chatter at your wife's next bridge party! One of its big advantages over that mechanical monster you're now using is quietness—but it's also easy to read, good looking, accurate and different.

**ham radio**

# low-power dummy load and rf wattmeter

**An accurate and reliable test instrument that can be built for less than five dollars**

Neil Johnson, W2OLU, 74 Pine Tree Road, Tappan, New York 10983

The advantage of using a dummy load for rf tests are many. No interference is created, and station identification isn't required for prolonged test periods. If the dummy load contains a calibrated meter, you can measure rf power with good accuracy.

Most hams have dummy loads that will handle a kilowatt. While these are fine for their intended purpose, most won't provide accurate readings at power levels of 25 watts or less.

The unit described in this article is a highly accurate dummy load and rf wattmeter that can be used for testing low-power transmitters. Two versions are described: one is limited to 11 watts, and the other is good for 25 watts. Essential components cost less than two dollars. The enclosure, connector, and handle bring the total cost to less than four dollars. Sound intersting?

## 11-watt load

My original load used eleven 2-watt resistors in parallel. To save a buck, I used ten compostion resistors, each 560 ohms, and one 680-ohm resistor. These are rated at 2 watts, 10 percent tolerance. I chose them because they're widely available and cost less in quantities of ten units.

The resistance of this dummy load is 51.73 ohms, which provides a good match for popular coaxial cables. Power capacity is limited to 11 watts. This is because composition resistors should not be used at more than 50 percent of their power rating, otherwise their resistance will change, which will degrade the accuracy of the wattmeter.

I mention this for those who might wish to build a very low-power unit with readily available resistors. The schematic of **fig. 1** can be used, substituting the

composition resistor bank for R (see photo).

## 25-watt load

Better resistors will allow rf power measurements up to 25 watts. The better units turned out to be another of those surplus bargains that appear from time-to-time.* I ordered two of these and connected them in parallel to obtain 50 ohms.

In an earlier project I used the 25-watt version of these resistors in a dummy load.[1] Four resistors, each 50 ohms, were connected in series-parallel. The CGW resistors can be extremely overloaded and will retain their accuracy to a remarkable degree. The 25-watt units used in the load described in reference 1 have been sold

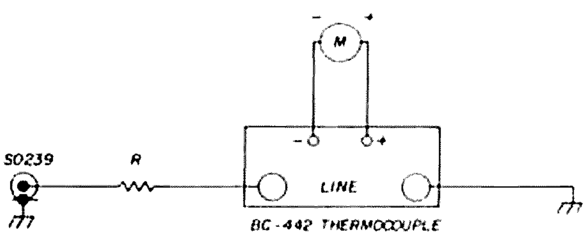



fig. 1. Schematic of the 25-watt dummy load and rf wattmeter. Resistor is two 100-ohm resistors in parallel; each resistor 13 watts, 1% non-inductive type.

out. However, for those who may have some, the data provided in table 1 should be helpful. I believe the overload characteristics with respect to permanent resistance change also apply to the 13-watt units.

## thermocouple and meter

These components are another "ham special" value.† Both units came from the BC-442 antenna tuning unit. They are first-rate, rugged, accurate components originally designed for military use. The dc meter has a nonlinear scale, a relatively expensive type of meter construction not

*John Meshna, Jr., 19 Allerton Street, Lynn, Massachusetts 01904. Catalog listing: "Corning Glass Works, Tin-Oxide Film Resistors; 100 ohms, 13 watts, 1% tolerance." Price: 35 cents each.

†Fair Radio Sales, P. O. Box 1105, Lima, Ohio 45802; $1.25 for both thermocouple and meter.

table 1. Data for the CGW 25-watt resistor.

| | |
|---|---|
| tolerance | 2.5 or 10% |
| stability | less than 1% permanent resistance change* |
| operating temp | 25°C |
| manufacturer | Corning Glass Works, Raleigh, N. C. |
| marking | CGW R35 25W 50 ohms |

*When operated at 10 times rated power for 5 seconds.

generally associated with amateur equipment. This gives an expanded scale at the low end, and rf output as low as 2 watts can be measured.

My experience with this meter and thermocouple has shown them to be highly accurate and dependable. This comes as no surprise when it's realized that these units had to withstand constant vibration in flight, hour after hour, plus the many shocks in taking off and landing at some rugged air strips.

## construction

Construction details are shown in the photos. The usual procedure of keeping rf leads short and direct should be followed in wiring this simple instrument. This will ensure a vswr of 1.1 or better throughout most of the amateur high-frequency bands.

Inside the wattmeter, showing the CGW "R Series" resistors.

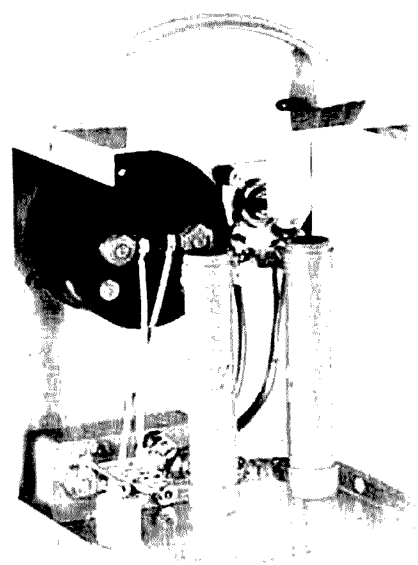


table 2. Calibration data.

| meter reading | true rf amperes | rf watts into 51 ohms |
|---|---|---|
| 1 | — | 0.9 |
| 2 | — | 1.8 |
| 3 | 0.23 | 2.7 |
| 4 | 0.28 | 4.0 |
| 5 | 0.335 | 5.6 |
| 6 | 0.39 | 7.9 |
| 7 | 0.46 | 11.0 |
| 8 | 0.54 | 15.2 |
| 9 | 0.64 | 21.4 |
| 10 | 0.78 | 31.4 |

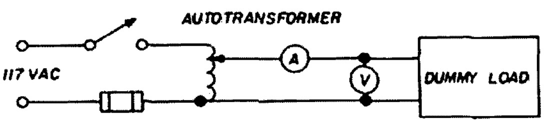



fig. 3. Circuit used for calibration. A is a Weston model 433 ac ammeter; V is a Simpson model 261 ac voltmeter (see text).

## calibration

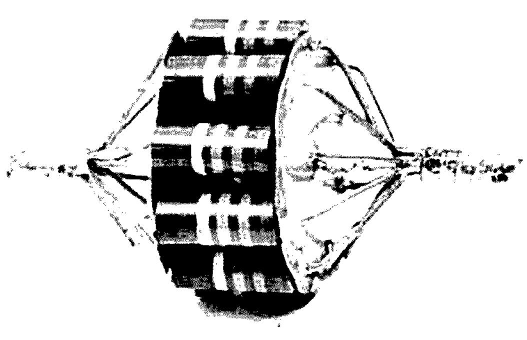
The calibration data shown in **table 2** and **fig. 2** were obtained with the setup shown in **fig. 3**. The wattmeter was calibrated at 60 Hz, using mirror-scale laboratory instruments: a Weston Model 433 ac ammeter and a Simpson Model 261 voltmeter.

The voltmeter takes power, so I checked it against the Weston 433 before starting the calibration procedure. Readings were checked going up and down the scale. Then a second check was made to ensure repeatability of the data.

Note that the meter reading at 10 (table 2) occurs when the resistors are operated beyond their maximum power rating. No harm is done to either meter or resistors. A temporary effect was that the resistors changed value to 50.5 ohms, and the entire dummy-load resistance increased to 51.5 ohms. The resistors returned to their original value upon removal of rf power.

The 51-ohm resistance of the unit is due to the series combination of the resistors (50 ohms) and the thermocouple, approximately 1 ohm. Note that readings below 0.23 ampere are not valid.

If you intend to use the instrument at levels above 15 watts for extended periods, some means of auxiliary cooling should be used. Removing the cover will also help.



Composition resistor bank used in the 11-watt unit.

## conclusion

The amount of abuse the CGW resistors will take staggers the imagination.

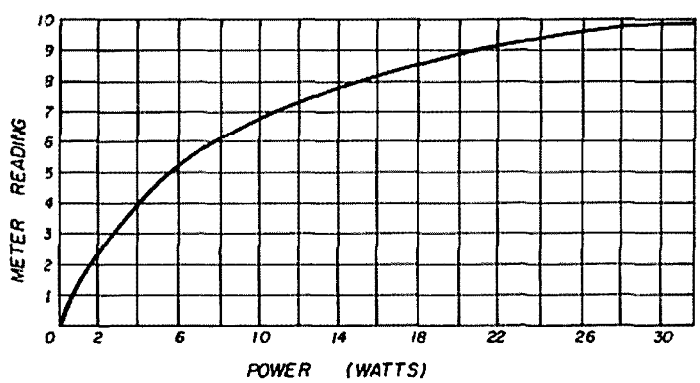



fig. 2. Calibration curve for the low-power wattmeter.

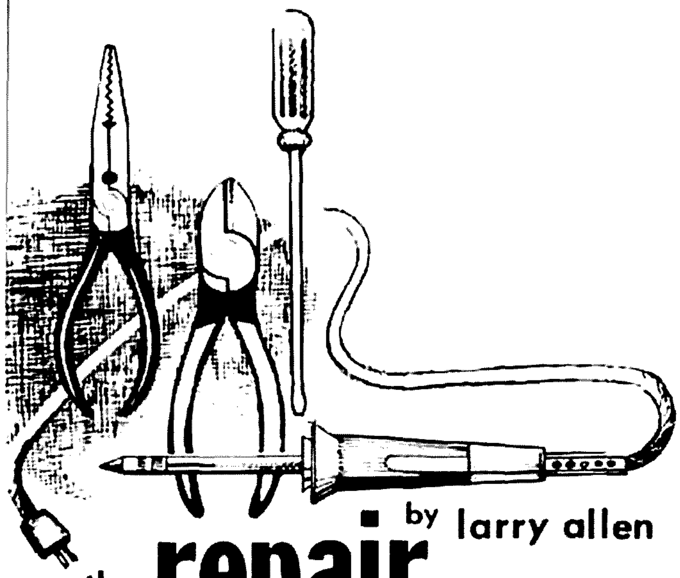
For example, my old 200-watt load[1] withstood a kilowatt for 5 seconds maximum without resistor values changing more than 1 percent (permanent change after cooling). By the same token, the smaller resistors, which have a nominal rating of 26 watts per pair, could be run up to 260 watts.

A high-quality rf wattmeter, good for any power up to 25 watts, is not a bad investment for $4 and a pleasant hour's work.

### reference

1Neil Johnson, W2OLU, "A 200-Watt Dummy Load for $2.00", *73* May, 1967, p. 66.

ham radio



■ by larry allen

the repair bench

# how to use a sweep generator

Television repair technicians – the good ones – learn to use a sweep generator early in their careers. It's a must. There's no other way to align a TV set, because its i-f stages are stagger-tuned. That is, they're aligned to various frequencies. It takes a sweep generator to show how they're tuned and the effect they have on the wideband television signal.

But I've discovered an awful lot of hams don't understand a sweep generator. If *you* don't, read on. I'm going to explain it.

You can use a sweep generator to help you design interstage coupling transformers and wind them with just the bandwidth you want. Or, if you tinker with ham TV, you'll need it to keep your receiver aligned. It's also a good trouble-shooting tool when your ssb receiver (or any receiver, for that matter) has rf or i-f trouble. You can even watch the effects of agc (or avc) on bandwidth.

## the instrument

Sweep generators for general use are scarce these days. Many new models are for television sweep alignment, and put only TV i-f and channel frequencies. But some general-coverage models are still available. Here's a list of those I've seen and used:

EICO 369
Heathkit IG-52
Knight-kit KG-687

The Knight-kit has a built-in marker generator; I'll explain its purpose, too, later.

You can see what they look like in **fig. 1**. These cover frequencies from about 3MHz to 220 MHz, the high end of the vhf TV band. You can often use harmonics of the dial frequencies up to nearly 900 MHz, with reduced output of course. Stability is usually critical at that high a harmonic. But the fundamentals, as you can see, encompass most ham frequencies.

What, exactly, is a sweep generator?

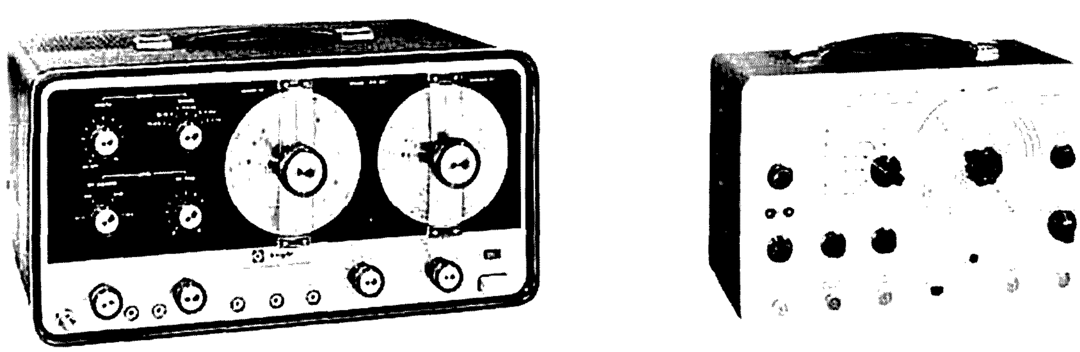
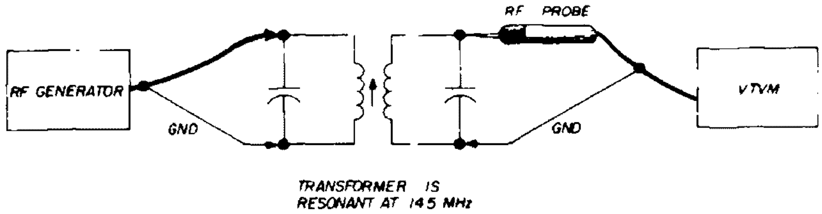
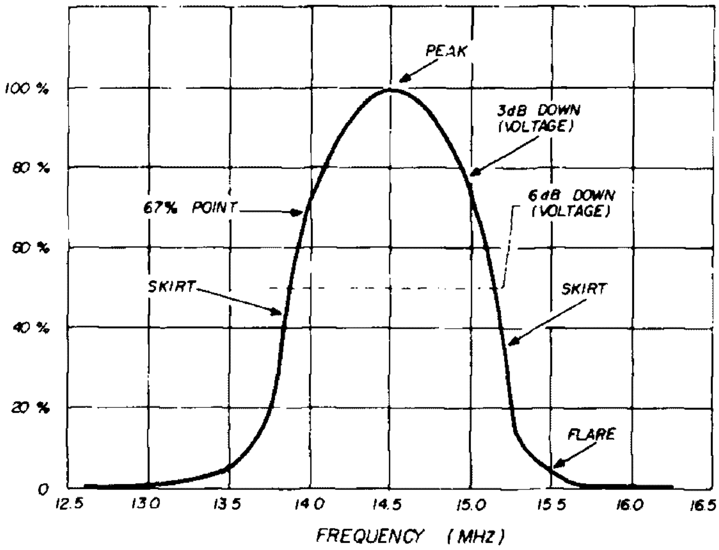
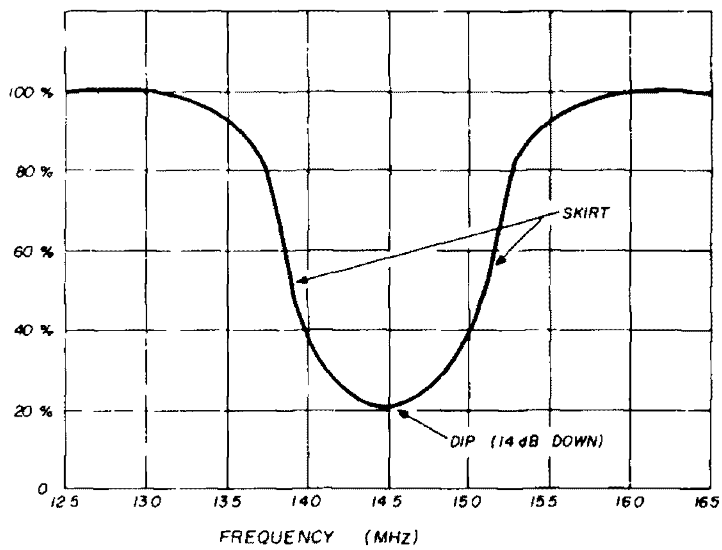


**fig. 1. Two sweep generators that are suitable for the amateur workbench.**

fig. 2. Hookup for measuring response by hand and graphs of parallel-tuned circuit (B) and series-tuned trap (C).




You'll understand better as I show you what it does, but here's a brief description.

An oscillator creates an rf signal at whatever frequency is set on the generator's dial. A sweep circuit in the generator is connected so it tunes the frequency of the main oscillator up and down, as if you were turning the dial above and below the frequency you first set it at. However, this sweep circuit is driven by a 60-Hz signal from the power line, and it swings the oscillator frequency up and down 60 times every second.

Suppose, for example, you set the dial for 14.5 MHz. The sweep circuit swings the oscillator above that and below it. If the generator is set to sweep 5 MHz, the oscillator is swung downward to 12 MHz, then up through 14.5 MHz to 17 MHz. It thus goes from center to one end, to the other end, and back to center, 60 times per second. The 5 MHz is called the *sweep width* and the 14.5 MHz is called the *center frequency*. The 60 Hz is called the *sweep rate;* that's the same on all service-type sweep generators.

## response curve

Now for some principles of what the instrument is for. Let me review tuned circuits. They are the only reason you ever need a sweep generator.

The most important characteristic of an LC circuit is the frequency it's resonant to. Second most important is how well it rejects nearby frequencies — in other words, how sharp its resonance is. Together, these are the *response* of a tuned circuit.

You can plot the response of any tuned coil-capacitor combination. You just feed in a lot of different rf signals, one at a time, and graph how the circuit responds to each one. The hookup is shown in **fig. 2A,** and the graph of results with one tuned circuit is **fig. 2B.**

Here's how it's done. Tune the generator to whatever frequency makes the highest reading on the vtvm. Mark that on the center of the graph at the 100% line, and write the frequency directly below it along the bottom of the graph. This one happens to be peaked at 14.5 MHz.

Then start slowly downward with the tuning dial of the generator. Above the 14.25-MHz line, make a dot that represents the voltage that gets through the

*transformer.* Show it as a percentage of what the voltage is at peak. As you tune the generator on downward, make voltage-level dots at every 0.25-MHz increment: at 14.0, 13.75, 13.5, 13.25, 13.0, 12.75, and 12.5 MHz.

broadband. The 6-dB points (half-way down) on its curve are at about 13.8 and 15.2 MHz. Its 6-dB bandwidth is approximately 1.4 MHz. At and below 13 MHz, response is nil; and it's nil at 16 MHz and above.

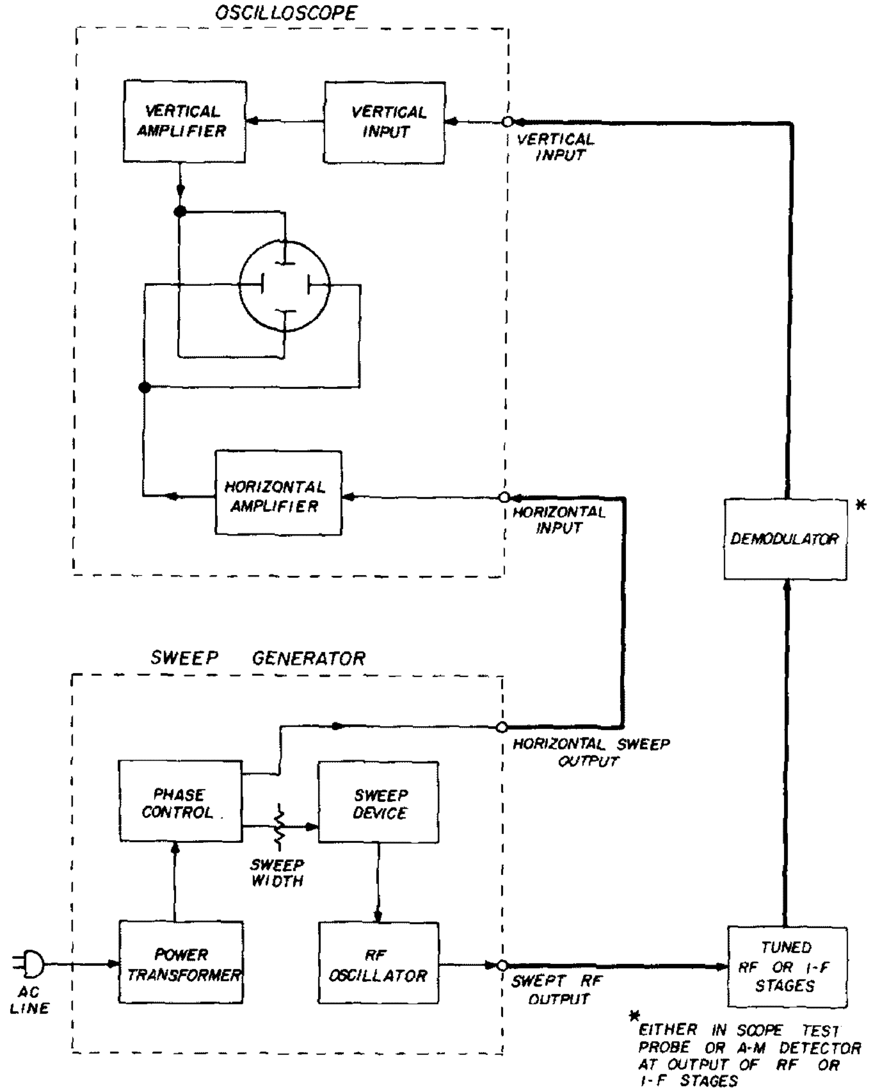



**fig. 3. Connections for displaying response curve on a scope.**

Then start again at the peak and move the generator dial upward. Mark the voltages at 14.75, 15.0, 15.25, 15.5, 15.75, 16.0, 16.25, and 16.5 MHz.

These dots represent how well the *tuned* transformer *responds* to each frequency. Joining the dots together with a solid line helps you interpolate responses at frequencies in between the increments you measured. The solid line forms a curve, which is a *response curve* for that particular tuned transformer.

This particular transformer is fairly

If you plot the response of a tuned circuit that's connected as a trap, the curve is inverted, as in **fig. 2C.** You plot this one starting at some frequency well below the expected response of the tuned circuit. The meter registers the full signal from the rf generator. As you turn the dial toward resonance, the rf voltage reaching the meter starts dwindling – at 13 MHz. It gets less and less. Once you tune past the resonant frequency, voltage of rf reaching the meter rises again. Eventually, you tune the rf generator

beyond the influence of the tuned circuit. That's beyond 16 MHz.

## plotting automatically

What if you could turn the generator dial back and forth very rapidly and just as rapidly plot *all* the voltage points instead of just some? You'd develop the response curve a lot quicker. Well, you can. In fact, the whole thing can be done automatically.

You use an oscilloscope as the voltage-measuring device. And you use a sweep generator to swing the frequency up and down rapidly – 60 times per second. If the scope is synchronized to move its beam back and forth 60 times each second, the voltages measured at all frequency points by its vertical amplifier are displayed side by side. The outcome is a continuous curve of voltages right on the screen of the scope.

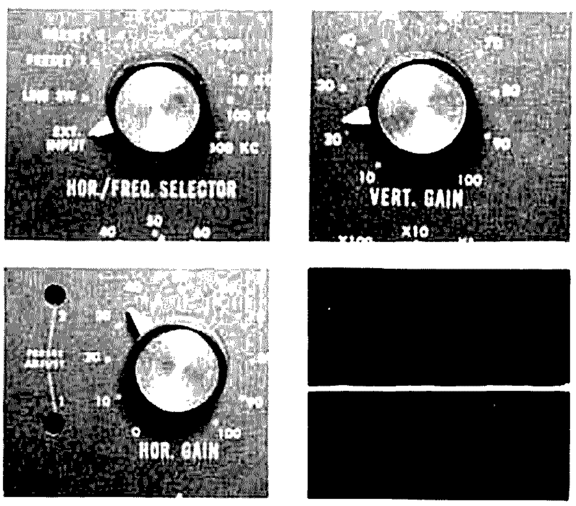



**fig. 4. Settings of scope controls to produce response curve. Photo at bottom right shows a baseline before the demodulated signal is applied.**

The sketch in fig. 3 lets you see in detail how to make the connections. The same 60-Hz voltage that operates the sweep circuit in the generator is fed by a connecting cable to the horizontal input of the scope. Applied to the horizontal deflection plates of the scope crt, it swings the beam back and forth exactly in rhythm with the sweeping of the rf oscillator in the sweep generator.

At the same time, the swept rf signal is fed to the tuned stages of circuits whose response you want to view. A demodulator probe (or a detector with the stages) develops an output voltage for each and every frequency in the band being swept, one right after another.

Those voltages go to the vertical plates of the crt. They move the beam upward in proportion to each voltage. Since the beam is at the same time being swept from side to side, each voltage level appears one after the other. And, since it is being swept exactly in step with the sweep-generator rf signal, the voltage levels occur in the same sequence as the frequencies.

This is done over and over, 60 times a second. Your eye sees an automatically plotted curve of the response. It's the response of the whole group of tuned circuits or stages to the band of frequencies being swept by the generator.

## sweeping a receiver i-f

As I said, there are dozens of ways you can use this ability to see the response of a tuned circuit or stage. So I'll show you how to set it up. The i-f of a ham receiver makes a good example, but this hookup can work for any tuned circuit or any group of tuned stages. Just feed swept rf into the input and feed the output to your scope.

First connect up the sweep generator and scope as diagramed in fig. 3. Horizontal output of generator to horizontal input of scope. Rf output of generator to input of tuned stages (in this example, to the input grid or base of the i-f section). Using direct probe, connect scope vertical input to output of a-m detector. (If there's no detector, use the scope's demodulator probe.)

The photos in fig. 4 show settings of the important scope controls. Horizontal Sweep to "Ext." Vertical Input attenuator to X1 (the object is to set up the scope so a 2-volt peak-to-peak input signal makes a 2-inch vertical display). Horizontal Gain up just enough to make a base line about 3 inches wide. Positioning

controls to keep base line below center of screen. (If the diode in your demodulator probe or the a-m detector is connected with its anode on the output side, position the trace above center, because the response curve will come out negative — below the base line.) The last photo of fig. 4 shows the scope crt screen, with the base line properly set up.

You set up the generator as pictured in fig. 5. Rf dial at frequency near center of bandpass you expect in tuned stages. Sweep Width about twice as wide as you expect. (The i-f in this receiver has a center of 5 MHz, and a normal bandwidth of 50 kHz. A very low sweep setting is called for — about 100 or 150 kHz. A television video i-f requires a sweep width setting of 10 or 12 MHz.) Phase control must be set after response curve is visible on scope crt. Output control set just high enough to produce 2-inch display on scope (exact setting depends on amplification in tuned stages being tested).

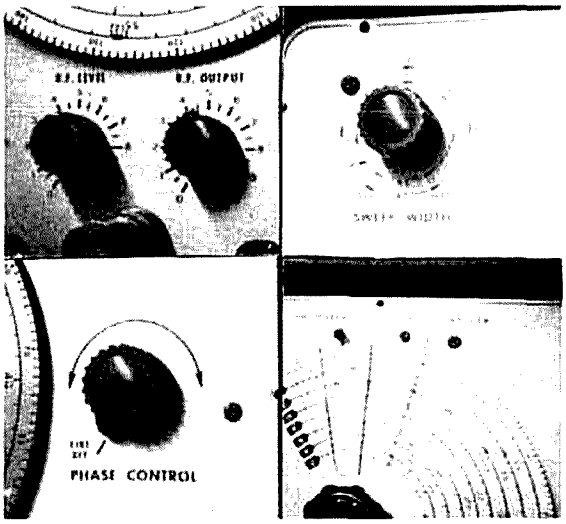
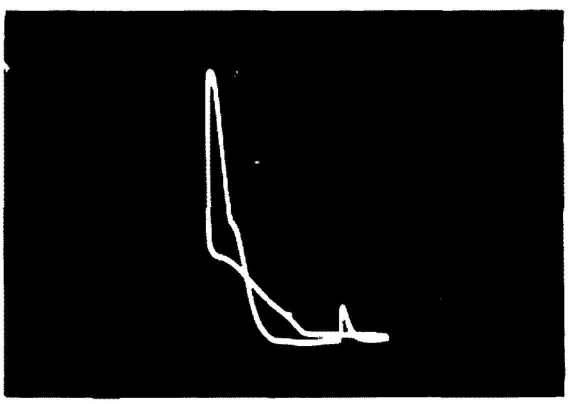
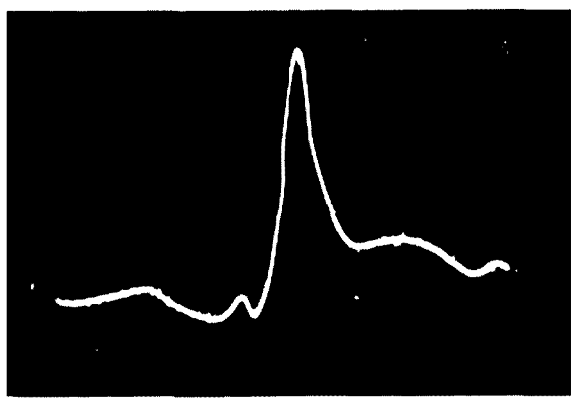


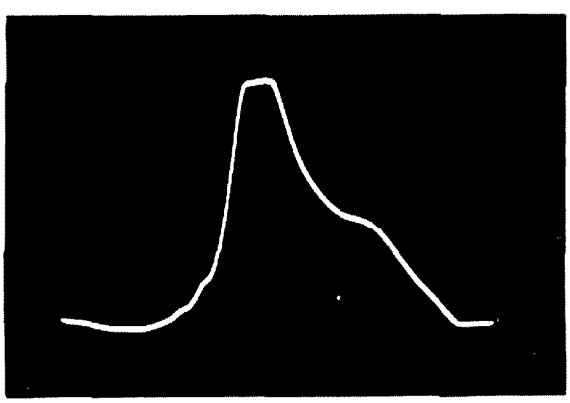
fig. 5. Control settings for the sweep generator.

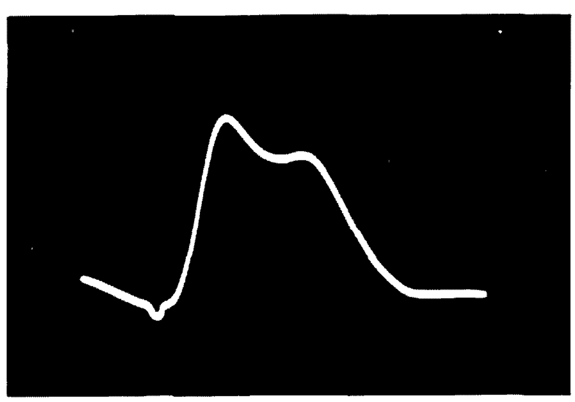
Fig. 6 is what response curves look like. The top left photo shows it before your adjust the Phase control on the sweep generator. The top right one is a normal response curve of an operating i-f section in a ham receiver. The curve of any tuned circuit, or any group of circuits tuned to the same frequency, should look

fig. 6. Response curves.


Phase control needs adjusting.


Normal narrowband curve.


Bandwidth spread out by misadjusted slug.


Wideband response of video i-f strip.

like this.

The bottom left photo is the response curve of the same i-f stages, but with two of the transformer adjustments misadjusted. As you can see, you can actually broaden out the response of the i-f, if that's what you want. For best QRM rejection, of course, you want the curve steep, narrow, and tall. That means, respectively, good adjacent-carrier rejection, good selectivity, and good amplification.

Incidentally, you might have to "clamp" the agc or avc line. Some receivers can't show a true response curve with the agc working. The alignment instructions will tell how much dc voltage to apply to the agc line if such a step is necessary.

## wideband tuning

The bottom right photo in **fig. 6** is the response curve of a television i-f strip. Any modern TV i-f has several stages tuned to different frequencies. It's called *staggered tuning*. For example, the input to the first stage may be tuned to 42.4 MHz, the interstage circuits to 43.0 and 44.5 MHz, and the output of the third stage to 45.0 MHz. These four response curves, when the stages are cascaded as TV i-f stages are, appear to line up side by side. The result is the wideband response curve you see.

There are trap circuits in a TV i-f, too. They are at 39.75, 41.25, and 47.25 MHz. Their major purpose is to eliminate any signals at those three frequencies; they interfere with the wanted signals. On the response curve, they are responsible for how steep the skirts are. Right beside the amplifying response curves, the trap curves drop the response faster at the edges than ordinary tuned circuits could.

## marking the frequency

One deficiency of a response curve displayed like this: you can't know exactly the frequency of the peak or of the points where response drops off rapidly. This is particularly important in a curve like the TV i-f response. You need to know where certain precise frequencies appear on the slope or along the top of the curve.

A *marker generator* and *marker adder* are the instruments for this purpose. At one time, they were separate instruments. Nowadays, the marker adder is part of either the sweep generator or the marker generator.

A marker generator is merely an accurate rf signal generator. Any stable rf generator can be used for marking re-

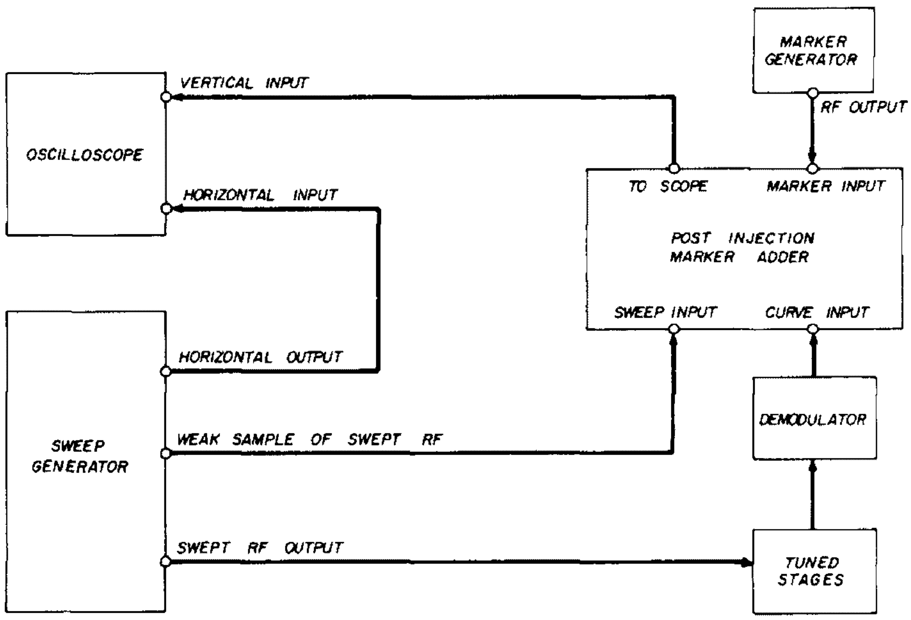



**fig. 7. Marker adder, shown external. Most modern sweep or marker generators have it inside.**

sponse curves, if it is accurate or easily calibrated.

The best way to use markers is by what's known as *post-injection.* The marker is added to the response curve after demodulation. (The old way was to feed the marker signal right in with the sweep-generator rf; it often upset the tuned circuits and made a false curve.) The sketch in **fig. 7** shows marker adder connections. The instrument is shown separately, but it's usually part of one of the generators. In that case, some of the connections are made internally.

For television, there are multimarker instruments available, with crystal-controlled marker signals. Such a generator displays markers at several points on the response curve simultaneously. The one at the left in **fig. 8** is an example. Marker frequencies are identified, to give you an idea how they help you recognize the true bandpass of the i-f stages being tested.

The ham-receiver curve in the middle has only one marker. It's labeled, so you know where (in frequency) the down-frequency skirt of the curve is. In actual practice, you find this frequency from the dial of the marker generator. The right-hand photo is the same curve, with a marker at the start of the up-frequency skirt. The frequency spread between the two skirts is the difference between the two marker frequencies. In this case, it's 0.1 MHz. That's the bandwidth of the i-f strip from which this curve is taken.

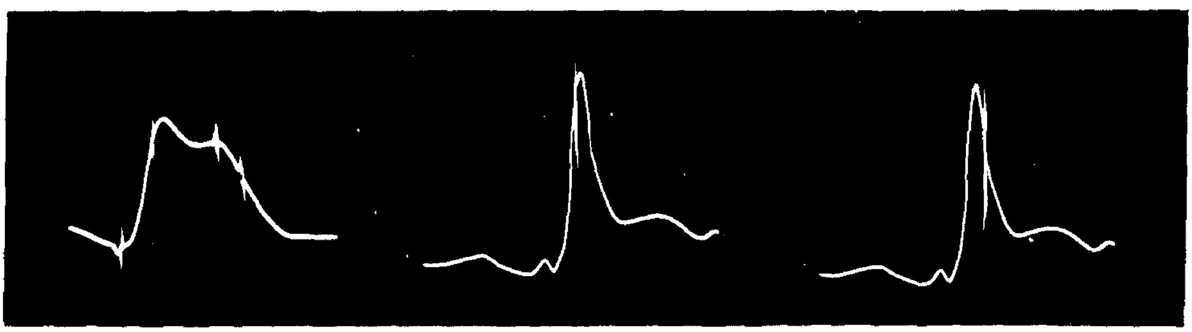


**fig. 8. Frequency markers on response curves. On the left the markers are provided with a special marker generator. Markers are used in the two photos to the right to show the bandwidth of the curve.**

## what's to come

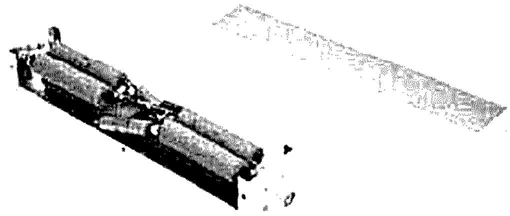
Armed with this information about how a sweep generator is used, you should find a lot of specific jobs for it on the ham bench. You can test and adjust filters, bandpass transformers, critical coupling between windings of transformers you wind yourself. With practice, you'll find the instrument easy to use. It tells you if your alignment job on a receiver (or the frequency conversion adjustment of an ssb transmitter) has produced the intended result. You can test any circuit that's within the rf range of the sweep generator you own.

Therein lies one drawback. None of the units available today go below a center frequency of 3 MHz. So, how do you do anything with 1.8-MHz circuits? Or the 60-, 455-, and 2.2-MHz i-f sections? If you find yourself using a sweep generator, and want to make it useful in this range of low frequencies, drop me a letter or postcard. If there's enough interest, I'll devote this department to that some month.

Next month, I write about mobile power supplies. They've come a long way since the days of the vibrator. Not every ham knows what to do with one when it's bad. Those transistor converter circuits look too complicated.

But they're not, really, Once you understand how and why they work, they seem simple. And fixing them on the repair bench is easy.

**ham radio**



# an all-band 10 dB power attenuator

This device can be used as a transmitter interstage buffer or for isolation when making antenna vswr adjustments

Chet Smith, K1CCL, 2 Jonathan Lane, Bedford, Massachusetts 01730

Occasionally a situation arises when it is preferable to provide interstage isolation with an attenuator rather than reduce drive to a following stage in a transmitter. For example, an ssb exciter may operate at maximum linearity at some level, say 100 watts. However, when used as a driving source for a transmitting converter, the exciter may cause distortion by overdriving the converter. Reducing input to the exciter will eliminate distortion in the converter at the expense of reintroducing it in the exciter. The buffer attenuator described in this article will resolve the problem, because it allows both units to operate at points of least distortion.

## other uses

Many amateurs use a simple reflectometer to measure vswr on the antenna transmission line. The reflectometer is reasonably accurate provided the source impedance doesn't change in response to varying load conditions. Most commercially built rigs and not a few homebrew ones are touchy about transmission-line vswr. When you're trying to match an antenna to such transmitters, a tip-off that something evil is at work is a change in apparent vswr when power level is changed. This is what happens when changing from a-m to cw on some rigs. A 10-dB pad between transmitter and reflectometer will correct this situation

by stabilizing the transmitter load. In addition, power level is reduced for matching purposes.

## circuit description

The circuit is a symmetrical double network (fig. 1). Its symmetrical configuration allows the attenuator to be inserted into the circuit without regard for "input" or "output" orientation.

A prototype circuit was derived from the equations given in reference 1 on the basis of a characteristic impedance, $Z_0$, of 50 ohms. The nearest values that could be obtained from components listed in the catalogs were used. The final version has a $Z_0$ of 55.6 ohms. Resistors were selected so that the attenuator would dissipate 100 watts in either open- or short-circuit condition.

An attenuation of 10 dB in the line will hold vswr to less than 1.25 regardless of the load impedance. This was achieved fairly closely in practice. It should be noted that the design uses resistors with nearest-available values and is **not** intended for precision applications.

The first unit (fig. 1A) used ordinary Sprague Koolohm[tm] power resistors. The inductive effects were quite noticeable on the higher-frequency bands. However, it was usable as a load buffer on 160 and 80 meters (vswr less than 1.5). Probably the reactance could have been compensated on 40 and 20 meters and higher bands.

Noninductive power resistors were used in the second unit (fig. 1B). This unit had an uncompensated vswr of less than 2.0 through 10 meters. The vswr increased to 2.56 on 6 meters (fig. 2). Although the resistors are rated as noninductive, there was some residual inductive reactance on 6 meters. A small capacitor was shunted across the network to tune out the inductive reactance. The vswr *then decreased* to 1.5 on 54 MHz, with no measurable change on 160 and 80 meters. Measured characteristics of the final circuit are presented in table 1.

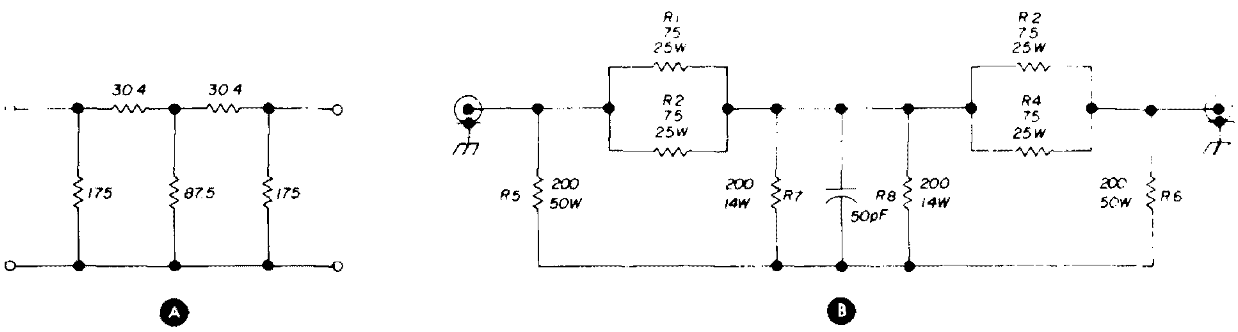



fig. 1. Attenuator schematic. A prototype circuit, A, was built with ordinary power resistors; $Z_0$ is 50 ohms. Final design, B, uses non-inductive resistors and has a $Z_0$ of 55.6 ohms. The small capacitor is needed to tune out residual inductive reactance.

| | |
|---|---|
| R1-R4 | 75 ohm, 25 W (Sprague 472E7505) |
| R5, R6 | 200 ohm, 50 W (Sprague 475E2015) |
| R7, R8 | 200 ohm, 14 W (Sprague 459E2015) |

## construction

The attenuator is housed in a Bud CU3013A enclosure; however any box measuring 1¾ x 2¼ x 10 inches will be satisfactory. Coaxial receptacles are type SO-239.

The attenuator is sensitive to the

table 1. *Measured characteristics of the 10 dB attenuator.*

| Band | Frequency (MHz) | vswr | excess phase delay* (deg) | insertion loss plus mismatch (dB) |
|---|---|---|---|---|
| 160 | 1.8 | 1.22 | 10.2 | 10.55 |
| 80 | 3.6 | 1.24 | 11.0 | — |
| 40 | 7 | 1.27 | 12.2 | 10.63 |
| 20 | 14 | 1.33 | 14.4 | — |
| 15 | 21 | 1.35 | 15.2 | 10.64 |
| 10 | 29 | 1.48 | 19.5 | 10.72 |
| 6 | 54 | 1.56 | 21.9 | 10.8 |

*Inductive; compensating capacitor could be larger.

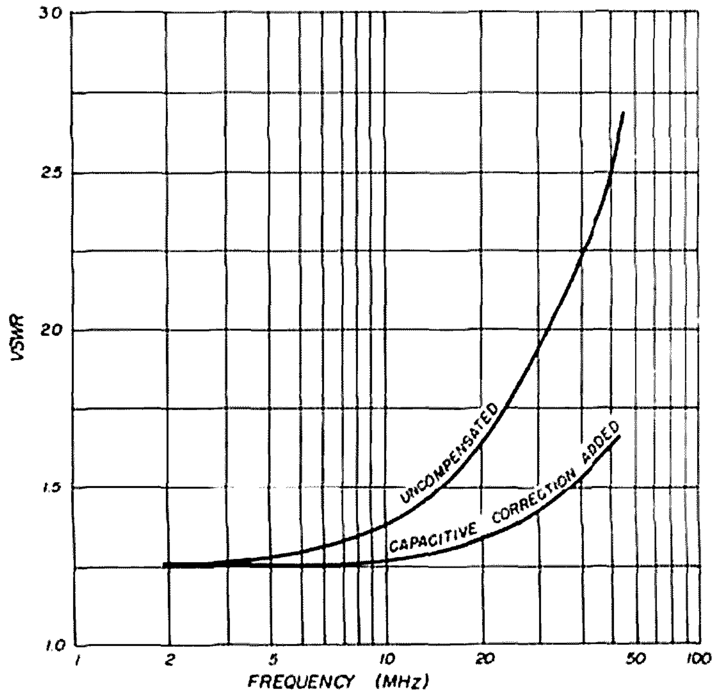



fig. 2. Effect on vswr with and without capacitive compensation.

immediate environment and could not be compensated with the cover removed. The unit was compensated originally with a small variable capacitor, which was replaced with a silver-mica fixed capacitor when the final capacitance value was determined.

## cooling

The attenuator was operated continuously with an input of 120 watts for over an hour. It was on a workbench, in the open air, and could be handled without discomfort. The package form factor has a high surface-to-volume ratio, which helps dissipate heat. However, if the unit is to be enclosed, a heat sink or forced-air cooling should be used.

A note of thanks is due to Bill Cobum, W1ELP, who built the attenuators and adjusted the compensating capacitor. Also, I'd like to acknowledge the help of Lee Tibbert, K1YOZ, who made the insertion loss, phase delay and vswr measurements.

### reference

1 *Reference Data for Radio Engineers*, FTC, 3rd edition, P. 158.

ham radio

# vswr alarm circuits

Here are some additions you can make to your vswr meter to give aural or visual warning of vswr changes

John J. Schultz, W2EEY, 40 Rossie Street, Mystic, Connecticut 06355

When you first put a new antenna into service, you're quite concerned about transmission line standing wave ratio (swr) and carefully note swr meter indications. During regular operation, however, you may glance occasionally at your swr meter, and if it's somewhere in the ball park, you probably don't bother to take an accurate reading.

A drastic increase in swr will cause abnormal transmitter tuning, so a moderate swr increase may not be noticed. But it's important to be aware of any increase in swr for several reasons. A high swr can cause component breakdown in high-power equipment as well as in low-power transmitters where marginally rated components are used. Also, a small but continuous increase in swr may indicate an incipient problem with your antenna or transmission line.

What you need is a circuit that will monitor swr and give an alarm when the swr exceeds a predetermined limit. This article discusses and illustrates a number of swr alarm circuits. All are intended to be used with in-line swr meters of the Monimatch type. In most cases the alarm circuit can be easily built into an swr meter enclosure.

Simple alarm circuits have been emphasized in the literature. You can combine some of these to form more complex alarm systems. Once the general requirements are understood, you may want to develop your own circuit. A large number of voltage- or current-level sensing circuits are shown in electronic circuit design handbooks. These can be adapted as swr alarms.

## basic principles

In-line swr meters use a diode circuit to provide a dc voltage proportional to the power in the transmission line in the

forward (incident) and reverse (reflected) directions. A switch allows either incident or reflected voltage to be read on the meter. The meter is calibrated to indicate swr in the reflected voltage position after the meter has been set to full-scale indication in the forward direction.

The loading effect of the meter resistance on the diode circuit reduces the voltage across the meter to a very low value. However, if the diode circuit is lightly loaded, several volts may be produced in the forward or reflected position that can be used to activate an alarm circuit (fig. 1).

With the swr meter switch in the forward position, an alarm circuit in the reflected-voltage diode output circuit can be set to operate when reflected voltage becomes excessive. If the alarm circuit's input resistance is reasonably high, it won't affect normal operation of the swr meter. When the meter is switched to the reflected voltage position, the low meter resistance won't be affected by the alarm circuit.

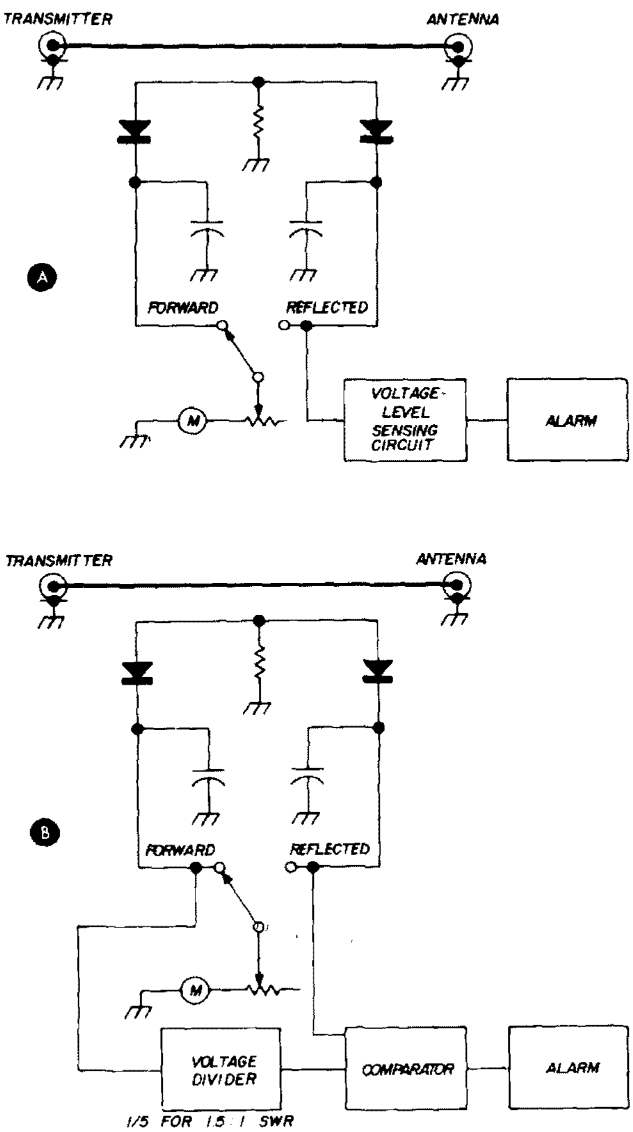



fig. 1. Methods for adding an alarm circuit to an existing swr bridge. Voltage-level sensing circuit, A, is simplest; comparator circuit, B, is effective over a wider range of operating conditions.

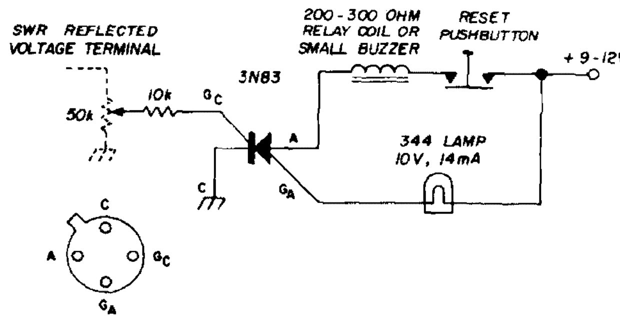



fig. 2. Simple scr latching-type alarm circuit requires only 0.4 to 0.9 volt trigger level.

## power and frequency considerations

This simple concept works well and is satisfactory for most situations. It does have some characteristics that should be considered, however. The diode circuit voltage corresponding to a certain swr remains correct only for the transmitter output for which it was originally established. If you usually operate a transmitter at the same power, then there's no problem. But if you use different power levels, the point at which the alarm circuit is activated has to be re-established for each new power level.

Another point to consider is the voltage pickup from the transmission line. It can vary with frequency depending on the pickup method. Some swr circuits will exhibit little variation in voltage pickup when used from 80 to 10 meters, while others will vary 60 dB or more. If there's a wide variation in voltage pickup, you can (a) set the circuit to enable at a narrow swr range on the higher-frequency bands, or (b) readjust the alarm for each band and accept the fact that a greater swr level will be required to activate the alarm on the lower-frequency bands.

## comparator circuit operation

If you operate only on one band with a transmitter of constant power level, the simple alarm circuit of fig. 1A will

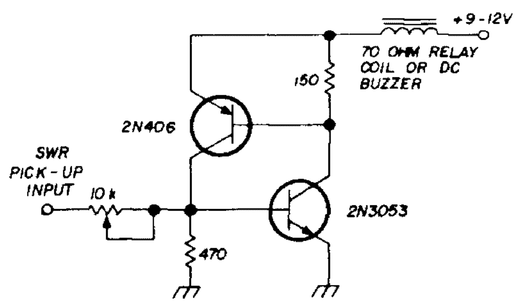



**fig. 3. Nonlatching voltage-sensing circuit. Relay will actuate when input voltage exceeds a preset level; 10k ohms potentiometer allows trip level to be set from about 0.4 to 10 volts.**

suffice. A slightly more complex approach can be used, however, that eliminates the problem of resetting the alarm for different power levels and different swr meter voltage pickup on different bands. Because the swr is determined by the ratio of reflected-to-forward voltage, you can use a circuit as shown in the block diagram of fig. 1B. Here, a voltage divider and voltage ratio sensing circuit (comparator) are combined to provide an alarm that operates irrespective of transmitter power.

Since this ratio is determined by the line swr only, it remains the same for any transmitter power level. The forward and reflected voltage amplitudes change the same with frequency, but their ratio does not. As shown in fig. 1B, you can preset the alarm for a specific swr level. For an swr of 1.5, the reflected voltage will be 20 percent of the forward voltage. Thus, if the forward voltage is divided by 1/5 and compared with the reflected voltage, the differences should never be greater than zero unless the chosen swr is exceeded. The comparator monitors the difference between the voltages and produces an alarm whenever a difference occurs.

## practical circuits

Practical swr alarms can be divided into meter-relay, voltage-level sensing and voltage comparator circuits.

The simplest alarm would be a meter-relay with an adjustable trip-point setting substituted for the regular swr meter movement. The relay circuit could be set to trip at some chosen swr level. Commercial meters of this type are available in ranges between 0.50 µA and 0-1 mA (Simpson 29XA series) but are quite expensive. If you can obtain such a meter at surplus prices, it will provide a very simple and reliable swr alarm. The relay contacts can't handle much current (about 10 mA), but they can be used directly to control an aural alarm device such as the Mallory Sonalert solid-state buzzers.

A simple voltage-sensing circuit is shown in fig. 2. A positive level of 0.4 to 0.9 volt at the gate of the SCR will fire it

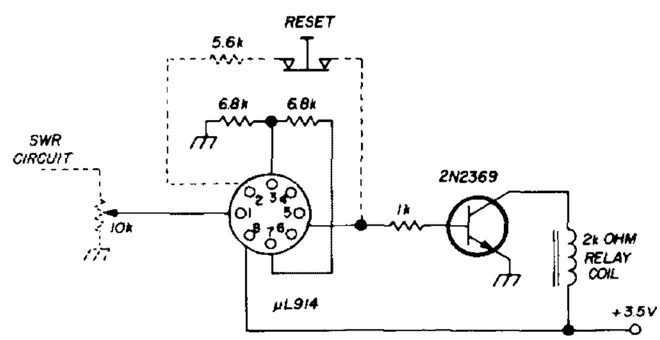



**fig. 4. Simple IC relay driver circuit provides positive trip action with input of about 1.3 V. Circuit can be made latching by including components shown in dashed lines.**

into conduction. Once fired, it will continue to conduct even if the gate voltage falls. Thus, a reset pushbutton switch is necessary in its anode circuit. The anode circuit can control a relay to activate some sort of alarm device, or a low-voltage dc buzzer can be driven directly. The lamp need not be used if it's not desired. The 10 kilohm resistor in the gate circuit provides very little loading on the swr meter circuit. The 50 kilohm potentiometer sets the reflected voltage level at which the scr will fire.

**Fig. 3** shows another voltage-sensing circuit that can be used in a manner

similar to the scr circuit. It's somewhat simpler because it won't lock or latch into a conducting state and, therefore, need not be reset. The transistors are inexpensive; each costs less than a dollar.

This circuit also provides a reasonably high shunting resistance to the swr meter circuit via the 10 kilohm input potentiometer that establishes the voltage level at which the relay coil (or buzzer) will be activated.

Fig. 4 shows still another voltage sensing circuit using an inexpensive μL914 integrated circuit. The circuit can be made either latching or nonlatching, depending upon whether or not terminals 2 and 6 of the IC are tied together.

In the foregoing circuits, the input voltage level that causes the circuits to operate is affected by the power-supply voltage. Essentially, no current is drawn from the power supply when the circuits are not conducting, so a simple battery supply can be used that will provide a constant voltage. If the voltage is obtained from a non-regulated source, a zener diode regulator should be used.

## comparator circuits

Voltage comparator circuits are shown in **figs. 5** and **6**. The circuit of **fig. 5** is complete with an audio oscillator included as the alarm device. One 1N34A diode is coupled to a 1/5 voltage divider (for a 1.5 swr) to the forward-voltage circuit of an swr meter. The other 1N34A diode is connected directly to the reflected-voltage circuit. If the reflected-voltage level becomes greater than the divided-down forward voltage level, the audio oscillator produces first a series of beeps. As the voltage difference increases, it produces a steady tone.

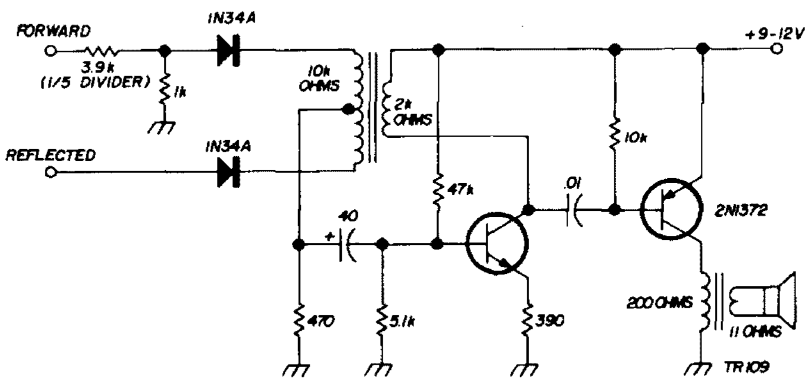



fig. 5. With this circuit an audio alarm is produced when a polarity difference exists between the input voltages.

**Fig. 6** shows a simple alarm using the Fairchild μA710 integrated circuit.* The IC comparator drives a 2N1711, which can be used to control an aural or visual alarm. (A pilot lamp in the 2N1711 collector circuit can be used for a visual indicator.) The μA710 inputs aren't shown as they would be connected to an swr meter; rather they indicate the basic application of the μA710. Its operation is as follows.

When the monitored voltage equals that of the reference voltage (at pin 3), the voltage at pin 7 goes from zero to about 2-3 volts. When the monitored voltage falls below the reference voltage, pin 7 returns to ground potential. Thus, for an application as an swr alarm circuit, the reference voltage can be obtained as shown in **fig. 5** (through a voltage divider from the swr meter forward-voltage circuit). The monitored voltage terminal would be connected to the reflected-voltage circuit.

The applications of the μA710 can be extended to many other alarm and control circuit uses. The only restriction is the maximum magnitude of the input voltages it can handle, which is about 7 volts.

*The μA711 Dual Comparator (essentially two μA710's in a TO-5 case) costs $2.50. Available from Poly Paks, Box 942, South Lynnfield, Massachusetts 01940, catalogue number 92CU813. Regular μA710's are $4.50 each.

## adjustment

Assuming the transmission line is matched at a low swr, a voltage-level sensing alarm can be connected to the swr meter reflected-voltage circuit, and the adjustment potentiometer can be set so the alarm is just activated. The potentiometer is then backed off slightly and locked. Any slight increase in swr should then activate the alarm. If you wish to set the alarm more exactly, dummy loads can be used to simulate the swr level at which the alarm will become activated. As mentioned before, this adjustment should be made at full power and reviewed on each

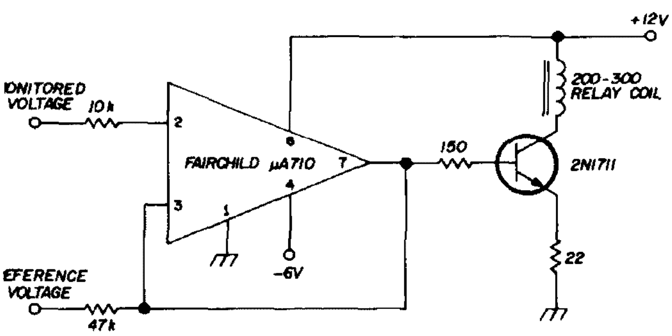



**fig. 6. IC comparator is ideally suited for swr monitoring but requires two supply voltages.**

band, because it may vary due to the swr meter pickup circuit characteristics.

The comparator circuits shouldn't require much adjustment. However, for exact setting, one leg of the voltage divider from the swr meter forward-voltage circuit should be variable. Then, using dummy load resistors at any reasonable power level to simulate various swr's, the potentiometer leg of the divider should be adjusted to activate the alarm at the desired swr. The loading effect of the voltage divider network may require a loading resistor in the reflected-voltage output circuit.

An added refinement I didn't explore is that of making the alarm self-powered by using the voltage in the forward pickup circuit. This may be practical for very low-current (lamp) alarms. Perhaps some readers might wish to develop this idea.

ham radio

# the ham notebook

## simple grid-block keying

In the May 1969 issue of *Florida Skip,* Dick Blasco, WA4DHU, describes a simple grid-block keying system that should be of interest to cw operators who build their own gear.

The merits of a differential grid-block keying system are discussed at length in many of the standard handbooks. However, most of the circuits described in the literature involve one or more vacuum tubes, many parts, and in general cause the builder to shy away from all that hardware and stick to a simpler, *poorer* system.

The system shown in fig. 1 is very simple, involves no active elements, and does a nice job of providing true differential keying at minimum cost. It is simply a voltage divider, with a zener diode providing the differential function normally provided by a voltage-regulator tube.

To design the system, determine the proper operating grid voltage and current to place your final amplifier in class-C operation. Resistors R1 and R2 are calculated as follows:

$$R1 = \frac{E_{gg}}{I_{g1}}$$

$$R2 = \frac{R1 \times E_{g1}}{2E_{gg} - E_{g1}} - 1000$$

where $E_{gg}$ is the negative bias voltage, $E_{g1}$ is the final grid bias and $I_{g1}$ is the final grid current.

The zener voltage is chosen by subtracting the vfo bias (key down) from the supply bias, and choosing a zener diode with a breakdown voltage smaller than this value. Several values should be tried to get optimum keying. C1 is chosen by trial and error to get the proper keying waveform, according to your particular taste. D1 is any general purpose diode.

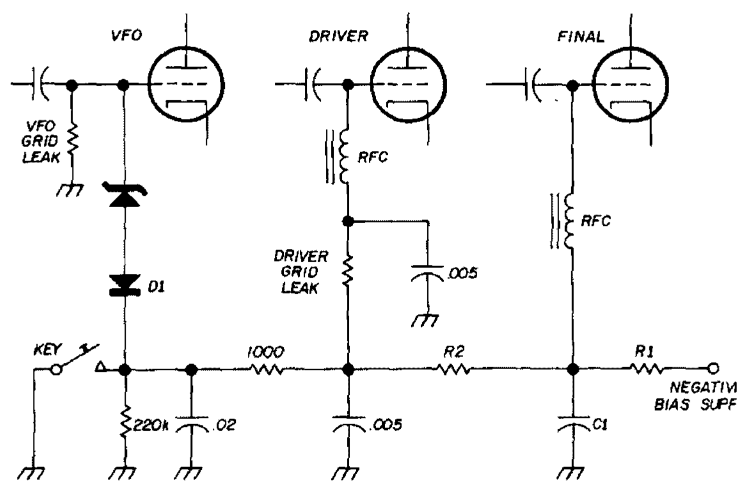


fig. 1. Simple grid-block keying system.

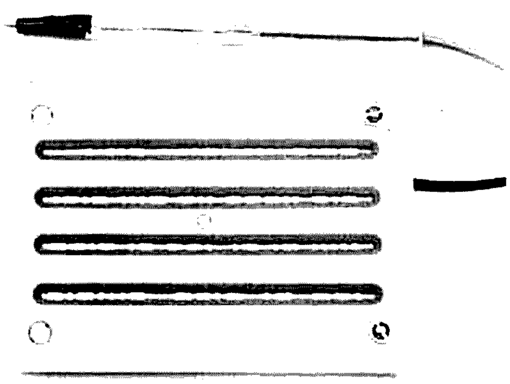
In operation, power amplifier bias falls exponentially as the key is closed. However, at a trigger point well before the driver and final conduct, the zener stops conducting and the vfo is turned full on, while the other stages become operative later (depending on capacitor C1). When the key is released the final and driver gradually turn off, but the vfo remains on until the bias is nearly at its peak value, at

which time the zener conducts, placing bias on the vfo to keep it from oscillating.

The circuit provides a minimum bias on the final of one-half the negative bias supply. This protects the final amplifier stage in case excitation is lost. Be sure that the grid leak in the vfo is at least 4700 ohms.

This circuit is a variation of the design used in the Heath HW-16 transceiver. Existing grid-block keying systems can be made differential by simply adding the two diodes to the vfo grid circuit.

## contest keyer



I built this gadget, the Scratch-1 programmable keyer, for the 1969 ARRL Field Day Contest. It was designed to replace an old keyer that used the rectified output of a tape recorder. However, the Scratch-1 offers several advantages: small size, no power connections, variable speed and access to different parts of the message.

Although the basic concept is quite simple, a number of prototypes were built to find an easy and durable keying system. The unit shown in the photo uses Scotch no. 49 aluminum tape for the keying element. As a preventative measure, a Q-Tip dipped in *NO-OX* (contact cleaner) was run over the keyer contact surface just before the field day contest. The keyer performed perfectly and shows little wear after completing over 300 contacts.

Since the code characters are fairly small, the smoothness of keying is critically dependent on how smoothly the operator can scratch the stylus down the base. I wanted a compact unit so I used quarter-inch dashes, but I would recommend making them a little longer to facilitate smooth operation. This keyer was designed for low-power applications such as grid-block keying – it probably won't take the rigors of cathode keying, particularly in medium-to high-powered transmitters.

Martin Davidoff, K2UBC

## mobile power supply

The simple mobile power supply shown in fig. 1, designed by WN8DJV, uses a low-cost toroidal transformer available from Fair Radio Sales . With no load, typical output is 540 Vdc; 500 Vdc at 50 mA; 480 Vdc at 100 mA; 450 Vdc at 200 mA; and 440 Vdc at 250 mA. Layout is not critical, but be sure you use a good, healthy heat sink for the two power transistors. If you have trouble finding a 3.3-ohm, 5-watt resistor, use three 10-ohm, 2-watt resistors in parallel.

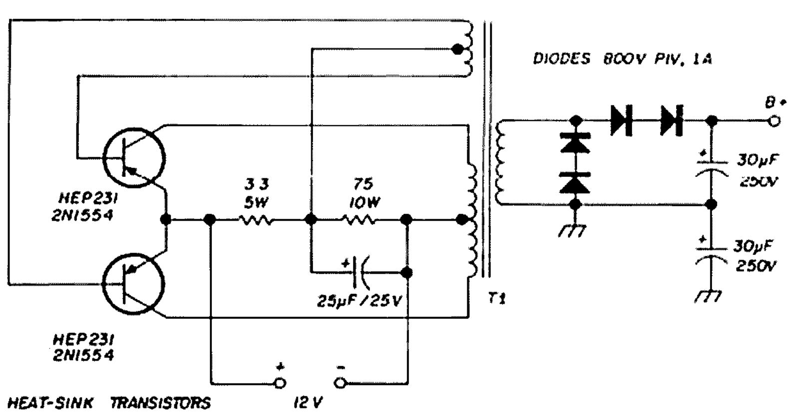
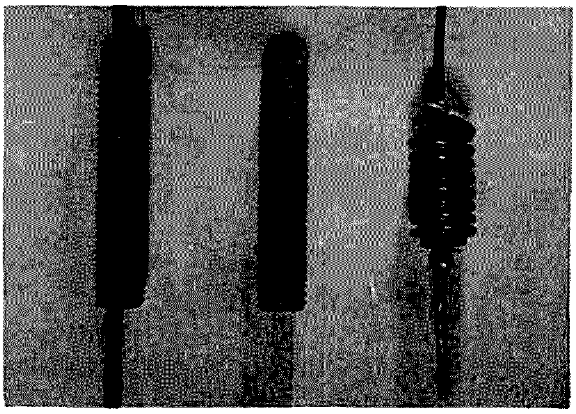



**fig. 2. Low-cost mobile power supply designed by WN8DJV uses an inexpensive surplus toroidal transformer available from Fair Radio. ($3.95, Fair Radio Sales, 1016 E. Eureka Street, Lima, Ohio 45802).**

## parasitic suppressor



Finding the right parasitic suppressor for a particular circuit can be a frustrating task. As a builder of a homebrew linear amplifier that had severe parasitic problems, I know what I'm talking about. I spent many days trying to suppress the parasitics and I finally came up with a coil-resistor combination that did the job on 80, 40 and 20 but wouldn't come through on 15 or 10. Since three out of five is better than nothing, I left the choke in. After operating for a few months and missing all the action on 15 and 10, I decided to try something else.

At the suggestion of a friend, I installed ferrite beads in the plate leads of my linear; I could not detect the slightest parasitic! Instant suppression. The magic of ferrite* does not stop here. It can be used for rf shielding and decoupling; grids can be shielded from strong rf fields, thus reducing instability.

The surprising thing about these little giants is that almost anyone with even a small junk box has some: they're used in coils and coil forms to increase inductance. The slug itself is about 3/16 inch in diameter and 1 to 1½ inches long.

When you slip a piece of wire through the slug, you create a low-Q inductor. The wire itself has a very small but finite *inductance* that is multiplied many times by the permeability factor of the slug, which is 900 or so. Although the inductance is still quite small, it's enough to suppress vhf parasitics.

**Jim Barcz, WA9JMY**

*Although the "beads" suggested by the author are probably powdered iron, and not ferrite, in this particular application the result is the same. Miniature ferrite beads are available from Amidon Associates, 12033 Otsego Street, North Hollywood, California 91607; package of 12 beads with a spec sheet and application notes is $2.00.

## rectifier terminal strip

Recently, while trying to find a suitable terminal strip for a bridge rectifier that used 8 diodes (2 in each leg), I decided to use an inverted octal socket. These sockets are fairly compact, and the diodes can be physically arranged similar to their schematic equivalent. This simplifies connections and eliminates costly wiring errors. One caution however: do not mount the socket directly to the chassis; use a bracket or standoffs since the contacts on some sockets can slide out the top of the socket and short the diodes to the chassis.

**Robert G. Wheaton, W5PKK**

## earth currents

Is the conversation in your evening QSO dragging? If so, here's something you can rig up in about five minutes that will open the door to all kinds of conversational possibilities. It is commonly known that the earth's magnetic field sets up currents that flow through the ground. What is not usually known is that these currents can be measured with the simplest of setups: a 50- or 100-μA meter, two pieces of brass welding rod and some wire.

Drive the rods into the ground 30 to 60 feet apart on an east-west line. Connect a wire to each. Hook the eastern rod to the positive meter terminal, the western rod to the negative. When the wires are first connected, the meter reading will be high because of galvanic action on the rods; after a few minutes the currents will drop to the normal value. If the meter reads down scale, install a reversing switch rather than changing the meter connections because earth current may reverse itself occasionally; or, use a 50-0-50 center-zero meter.

Different parts of the country have widely varying readings. It has been noted

at several installations near my home that erratic and fluctuating meter readings seem to precede an aurora opening on 6 or 2 meters. It is interesting to note that weather also has some effect on the reading: if the ground is wet the readings will usually be slightly higher. Magnetic storms have a noticeable effect.

One important note; be sure to put a .01 μF bypass capacitor across the meter terminals to eliminate the effects of rf or stray 60 Hz current.

**Don Samuelson, W7OUI**

## coaxial-line resonators

Transmission lines made of coax, twin-lead or "plumbing" are often used in vhf, uhf and microwave equipment. They can be used as (or in) filters, as tuning elements in amplifiers or oscillators, and as replacements for inductors or capacitors in tuned circuits. At frequencies over 100 MHz or so, transmission lines can be of practical lengths, and they exhibit high efficiencies and excellent performance and reproducibility, unlike some circuits using conventional lumped capacitors and inductors.

Unfortunately, it is sometimes hard to remember which type of transmission line exhibits which characteristics. To help out, table 1 summarizes the properties of shorted and open lines in terms of electrical length. The table lists wave lengths only between zero and $\lambda/2$, but the addition of $\lambda/2$, or any multiple of $\lambda/2$, does not change the properties.

The electrical length, $l$, of a transmission line is easily found:

$$l = \frac{l_p}{V}$$

Where $l_p$ is the physical length in meters, and V is the velocity factor (1 for air, 2/3 for common coax, 4/5 for small 300-ohm twinlead). Or for lengths in feet:

$$l = \frac{3.28\, l_p}{V}$$

The actual wavelength in meters is also easily determined:

$$\lambda = \frac{300}{f}$$

Where f is the frequency in megahertz. For example, a coax line shorted at one end is 1 meter long. What is its electrical length at 6 meters? Since V is 2/3 for common coax,

$$l = \frac{l_p}{V} = \frac{1}{2/3} = 3/2$$

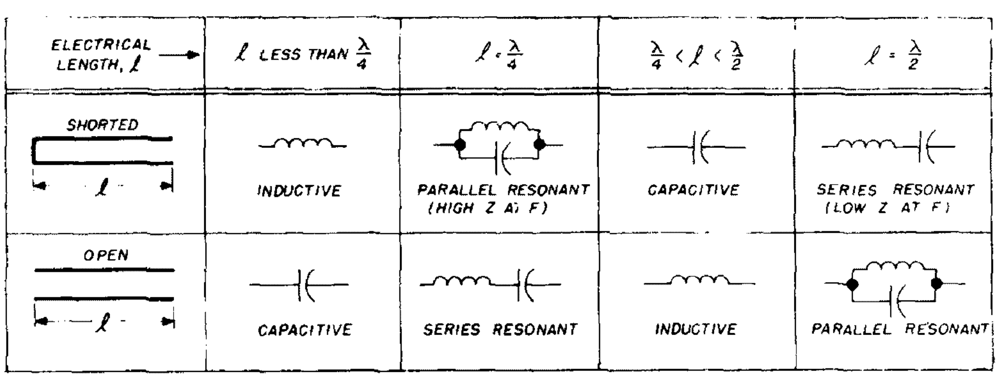



fig. 3. Electrical properties of transmission-line resonators.

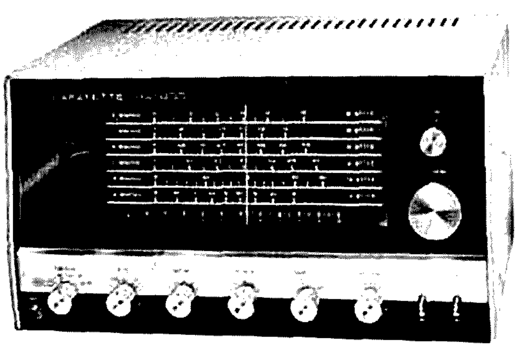
The electrical length is 1.5M, so is one-fourth wavelength ( $\lambda/4$) and acts as a parallel-tuned resonant circuit.

The transmission line can be loaded with a capacitor for tuning or to shorten the necessary line length. The effect of the capacitance depends on line impedance, location of the capacitor, frequency, etc. Either experimentation or consultation of a reference containing details will help out here.

**Paul Franson, WA7KRE**

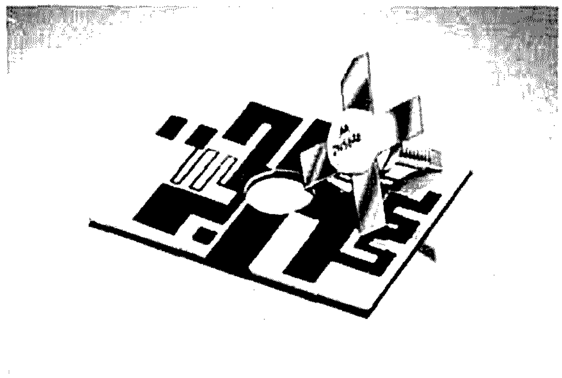
# new products

## six-band receiver



Lafayette Radio has announced a new solid-state six-band amateur-band receiver that incorporates three field-effect transistors and two mechanical filters to assure high selectivity with superior rf overload and noise suppression. The built-in power supply permits operation on either 117 Vac or 12 Vdc. Sensitivity is better than 1 μV on 80, 40 and 20 meters, 0.5 μV on 10 and 15 meters, and 2.5 μV on 6 meters. The double-conversion design uses a first i-f at 2.608 MHz and second i-f at 455 kHz; image rejection is better than 40 dB. Audio output impedance, 8 and 500 ohms; audio power, 1 watt; antenna input impedance, 50 ohms. $149.95. For more information, write to Lafayette Radio Electronics, 111 Jericho Turnpike, Syosset, L. I., New York 11791.

## rf power transistors



Motorola Semiconductor has announced a new line of balanced-emitter rf power transistors that can provide 12 watts at 450 MHz with a 12.5-volt power supply. The new transistors, the 2N5644 through 2N5646, are intended for use as power amplifiers for the commercial uhf mobile fm band, but are also suitable for amateur applications on 144,220 and 432 MHz. Since these devices are specifically designed for operation from 12.5 volts, no dc-to-dc converter is required.

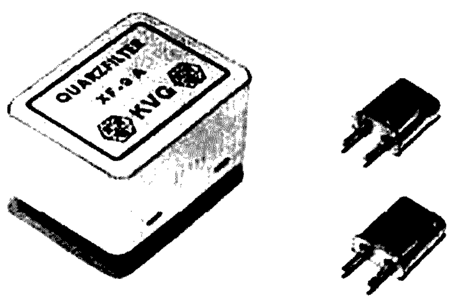
The 2N5646 family is supplied in the popular 3/8-inch ceramic stripline opposed-emitter package. Low-inductance leads provide easy design and adjustment, especially in broadband circuitry. An important feature of the new transistors is their balanced-emitter construction: each device is composed of many monolithic transistors in parallel with a nichrome resistor in series with each emitter. If emitter current tends to increase in any one of the transistors, the rise in voltage across the emitter resistor decreases base-emitter voltage, reducing the current flow. The equivalent resistance of all the resistors is very low, so they don't cause significant degeneration. Because of the balanced-emitter construction these transistors are very resistant to damage from mismatched loads or detuning. For more information, contact the Technical Information Center, Motorola Semiconductor Products, Inc., Box 20924, Phoenix, Arizona 85036. A similar line for 25 watts output is under development.

## rf amplifier ic

Motorola's new MC1590 rf/i-f amplifier features more than 20 dB increase over previous ic amplifiers in rf power gain (45 dB typical at 60 MHz) and agc capability (60 dB minimum from dc to 60 MHz). The high gain and wide-range agc are especially useful in portable receivers where wide ranges of signal levels are encountered. The wide-range agc allows the device to be used in the audio range as a speech compressor; it can also operate as a general-purpose amplifier up to 150 MHz.

The agc function has negligible effect on the i-f passband response because the input and output impedances of the device remain essentially constant during agc operation. For more information on the new MC1590, write to Technical Information Center, Motorola Semiconductor Products, Inc., Box 20924, Phoenix, Arizona 85036.

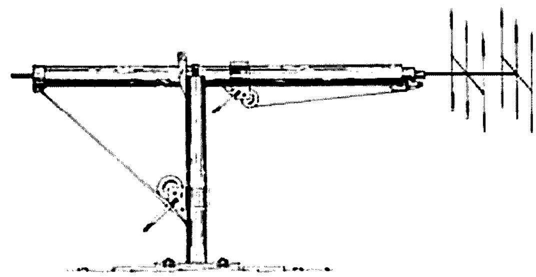
## Kvg crystal filters




The KVG company of West Germany is a major supplier of amateur and commercial crystals and crystal filters in Western Europe; you see them used in much of the amateur radio equipment built and advertised in the European amateur magazines. The KVG line of high-quality high-performance 9-MHz crystal filters and matched oscillator crystals for ssb, a-m and cw are now offered in the United States and Canada through Spectrum International.

Currently, six filter models are available: the XF-9A 2.5-kHz bandwidth filter for ssb transmitters ($21.95), the XF-9B 2.4 kHz bandwidth filter for ssb receivers and transceivers ($30.25), the XF-9C 3.75 kHz and XF-9D 5.0 kHz bandwidth filters for a-m ($32.45), the XF-9E 12.0 kHz filter for fm ($32.45) and the XF-9M 0.5 kHz bandwidth filter for cw ($23.00). Matching miniature HC25/U oscillator crystals for ssb operation (with the XF-9A and -9B filters), and cw heterodyne as well as 9.0 MHz a-m/cw carrier are $2.75 each and supplied complete with socket and socket fastener. For complete specifications on these crystals and crystal filters write to Spectrum International, P. O.Box 87, Topsfield, Massachusetts 01983.

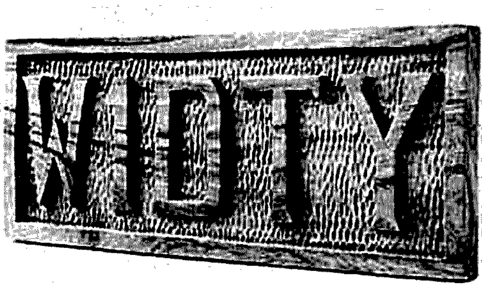
## crank-up antenna masts



The Tristao Tower Company has introduced a new series of self-supporting crank-up masts that are built from telescoping sections of high strength tubing. The *magna mast* – the larger of the two series – features self-supporting construction with rotor bases, cabling for uniform raising of section, geared raising winches with automatic brakes for maximum safety and one-man installation with a swing-over design that permits antenna servicing at ground level (no ladder or service platform required). The *magna mast* comes in two different sizes: 49 feet and 66 feet. Each is designed to support 12 square feet of antenna in 60 mph winds.

A smaller version of the self-supporting rotating masts is called the mini mast and is available in 30- or 35-foot heights. This mast will support 6 square feet of antenna in 50 mph winds. For more information on these new self-supporting crank-up masts, write to the Tristao Tower Company, P.O. Box 115, Hanford, California 93230, or Olympic General Corporation, P.O. Box 64398, Los Angeles, California 90064.

## wall plaques



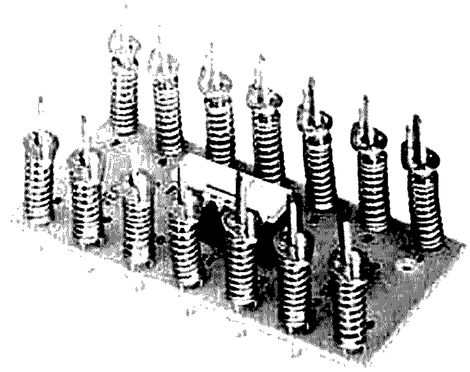

If you want to dress up your operating room, or if you're looking for a gift for another amateur, a hand-carved monkey-pod call-letter plaque like that shown above is an ideal solution. These handsome plaques are handmade in the Philippine Islands and are available with 4, 5 or 6 letters. The letters are about 5 inches high and the plaques are 8 inches high, 20 inches long and nearly 1 inch thick (five-letter plaque). The four-letter plaques are 17 inches long; the six-letter plaques, 22 inches. The price of the four-letter plaque is $11.30 five-letter, $11.90 and six-letter, $12.50 This includes Parcel Post shipment to the United States.

To order your plaque, send your check or money order to W. J. Chapman, W6DQE, 10208 Roscoe Boulevard, Sun Valley, California 91352. He forwards the orders to DU1FH via air mail; when the plaque is completed it is mailed directly to you from the Philippines. Delivery is about 2½ months after the order is received.

## technical aid

The amateur radio club at Sams Technical Institute (STI) has undertaken a "helping hand" program to assist amateurs and would-be amateurs with technical problems. This service is available to hams in all parts of the world, and is aimed particularly at those who are isolated from other amateurs or are unable to get local advice. Members of the STI Amateur Radio Club are students working toward an Associate Degree in Electronics Engineering Technology, and have a well-trained staff of instructors to back them up. If you have a technical problem that you are unable to solve, write to the STI Amateur Radio Club, WB9ADF, Sams Technical Institute, Interstate Industrial Park, Fort Wayne, Indiana 46808. STI is affiliated with ITT Educational Services, Inc.

## ic breadboard socket



A new device is now available for breadboarding with 14-pin dual-inline-package integrated circuits. The device consists of a small piece of epoxy-glass board with a 14-pin socket and two adjacent rows of Vector Springclip™ terminals; it is furnished with two pins on the bottom that may be press-fit mounted on prepunched Vectorboard (pattern AA).

This new gadget speeds up integrated circuit breadboarding because as many as four solderless connections can be made quickly with ordinary hookup wire. Logic circuitry is particularly easy to set up with the new socket because the numbered terminal pins correspond to socket contacts and interconnections between sockets can be made quickly and easily. If discrete components are required, they can be mounted on the adjacent Vectorboard.

The price of the 570G breadboard socket is $3.95, and may be ordered from Vector Electronic Company, Inc., 12460 Gladstone Avenue, Sylmar, California 91342.

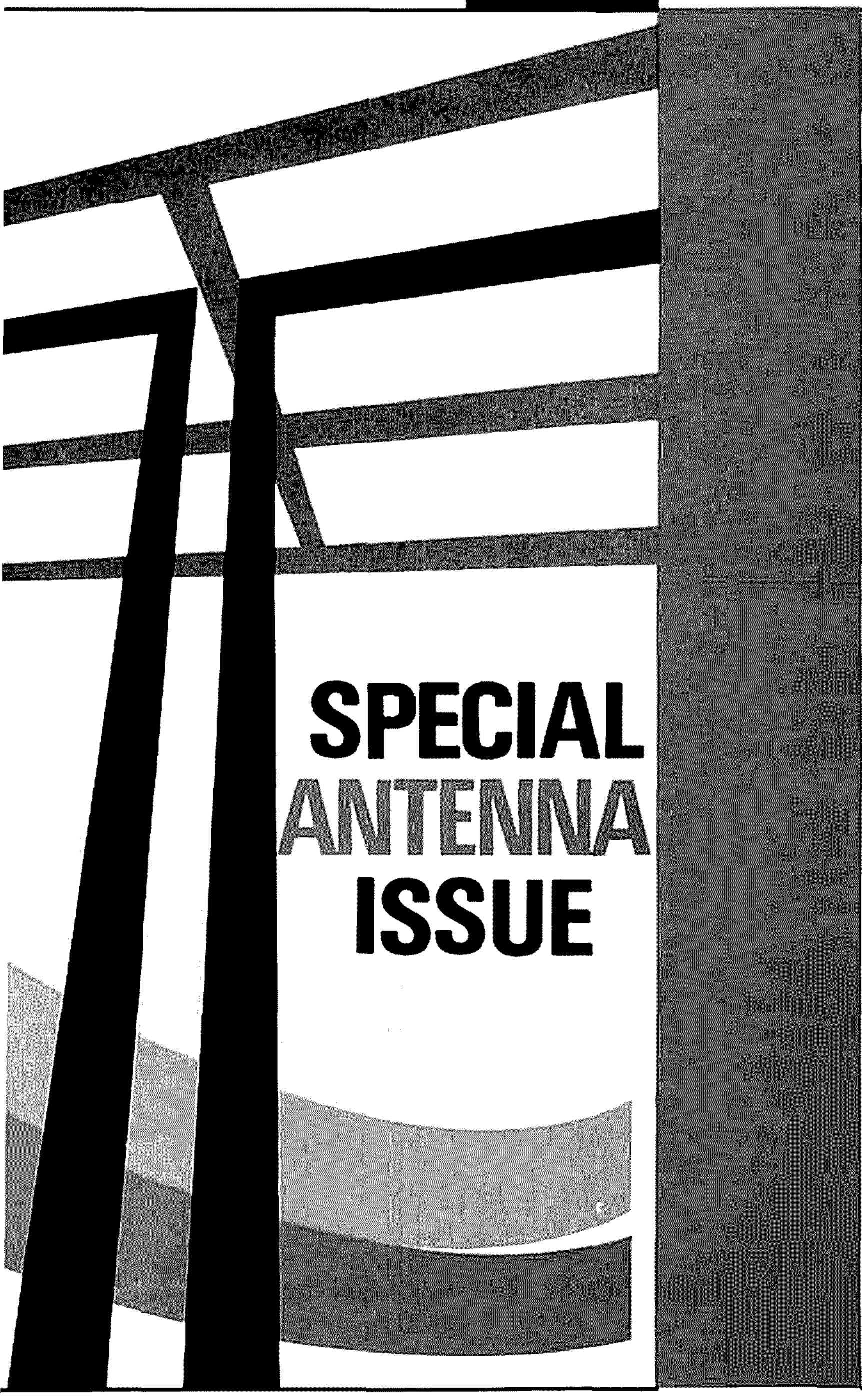
75 cents

focus on communications technology . . .

# ham radio

magazine

MAY 1970




## this month

may 1970
volume 3, number 5



# ham radio magazine

# contents



**offices**

Greenville, New Hampshire 03048
Telephone: 603-878-1441

**ham radio** magazine is
published monthly by
Communications Technology Inc.
Greenville, New Hampshire 03048

Subscription rates, world wide:
one year, $6.00, three years, $12.00.
Second class postage
paid at Greenville, N. H. 03048
and at additional mailing offices

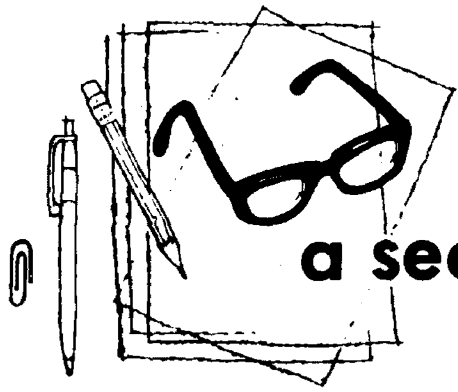
Foreign subscription agents:
United Kingdom:
Radio Society of Great Britain,
35 Doughty Street, London WC1, England.

All European countries:
Eskil Persson, SM5CJP, Frotunagrand 1,
19400 Upplands Vasby, Sweden.


Printed by Wellesley Press, Inc.
Wellesley, Massachusetts 02181, U.S.A.

**ham radio** is available to the blind
and physically handicapped on magnetic tape
from Science for the Blind,
221 Rock Hill Road, Bala Cynwyd,
Pennsylvania 19440.
Microfilm copies of current
and back issues are available
from University Microfilms,
Ann Arbor, Michigan 48103.

*Postmaster: Please send form 3579 to*
**ham radio** magazine, Greenville,
New Hampshire
03048



# a second look

by jim fisk

If you're interested in amateur vhf fm, you've probably already heard about the new FCC docket that details proposed rulemaking concerning amateur repeaters. The docket is the result of several rule-change proposals which have been submitted by interested individuals and organizations. The FCC commissioners read these "free lance" proposals, add other information they may have at their disposal and prepare a docket which reflects what *they* feel best suits the needs of the amateur service.

The new docket, Docket 18803, has drawn strong criticism from repeater enthusiasts who feel that the restrictions it proposes would serve to make it practically impossible to legally continue operating vhf fm repeaters! The proposed frequency changes—if the docket becomes law—alone would cost amateur vhf fm enthusiasts hundreds of thousands of dollars in new crystals! Never before has the Commission tried to dictate frequencies as outlined in Docket 18803—I wonder if this portends things to come on other amateur bands?

In addition to the receiving and transmitting sub-bands proposed in Docket 18803, it is proposed that the repeater log include only the time and date of the periods the repeater is available for service, as well as entries indicating the technical and operation condition of the repeater. It is also proposed to amend the present identification rule to permit automatic repeater identification by *CW* at intervals not to exceed three minutes.

Furthermore, Docket 18803 proposes that a repeater be designed so that it will normally be activated only by means of a coded signal or other means which will effectively exclude transmissions by stations not desiring to work through the repeater, thus minimizing unnecessary transmissions and the possible resulting interference (the coded signal may consist of a single audio tone so the repeater can easily be "whistled on"). Possible interference is also cited as the reason for outlawing simultaneous re-transmissions on two or more bands and cross-band operation. The proposed receiving and transmitting sub-bands are obviously a result of the same thinking.

However, the sub-bands tend to subject the repeater user to QRM problems while purporting to protect an unknown amateur not interested in repeater operation. How unnecessary, and unworkable, especially on the 420-MHz band. The specified channel spacing is too close for most of the equipment available to amateurs. Furthermore, when the present commercial equipment is surplus and available to amateurs, some of it will be useless because it is designed to work

(continued on page 75)

# unusual cubical-quad antennas

Four novel and interesting quad designs from overseas

Jim Fisk, W1DTY

**For all practical purposes,** high-performance antennas for 10, 15 and 20 meters fall into two categories: Yagis and cubical quads. Each has its pros and cons, its advantages and disadvantages, but the quad has probably been the subject of more amateur experimentation than the Yagi. The reason is simple: the quad is used primarily on the amateur bands while the Yagi sees service in commercial and military applications as well.

The cubical quad offers surprisingly high gain and excellent front-to-back ratio with a relatively short boom. Couple this with a low-angle vertical radiation pattern and you have performance that's hard to beat. However, the quad is not without its disadvantages: it's large, ungainly and offers high wind resistance. Fiberglass spreaders have solved some of the mechanical problems, but bamboo poles are still used by many, particularly overseas. Some all-metal quads have been built, but this requires careful design because of the large mass of metal in the antenna's rf field.

The antennas described below are variations on the basic quad theme - antennas that were designed to overcome some of the mechanical problems of building it, putting it up and keeping it there. These are quads that were designed with different feed systems, new supporting structures, different construction approaches—interesting cubical quad designs that have received little publicity outside the country where they originated.

## english quad

One of the first variations of the basic cubical quad was G4ZU's bird-cage antenna shown in **fig. 1**. This antenna, popular in Europe a few years ago, received little notice from the American amateurs. However, it is an interesting design that solves some of the electrical and mechanical problems associated with the true quad.

Although the bird-cage uses two full-wave loops, the upper and lower sides are made from aluminum tubing bent into a "V." This simplifies the supporting

structure, lowers the wind resistance and puts the low-impedance feed point near the mast. The vertical elements joining the horizontal aluminum supports are made from wire. Since the points where the antenna is attached to the vertical rotating mast are points of minimum rf voltage, insulation is simplified and weather losses are minimized. (Low-cost insulating material suggested by G4ZU includes wax-impregnated wood and bakelite.)

The stubs in the horizontal members of the antenna are used for tuning the system to frequency. It has been suggested that by increasing the vertical wires to somewhat more than ¼ wave, the antenna would resonate below the desired band. Variable capacitors could be inserted in each loop near the feed point to tune the antenna (the radiator being tuned for minimum swr, the reflector tuned for maximum forward gain).

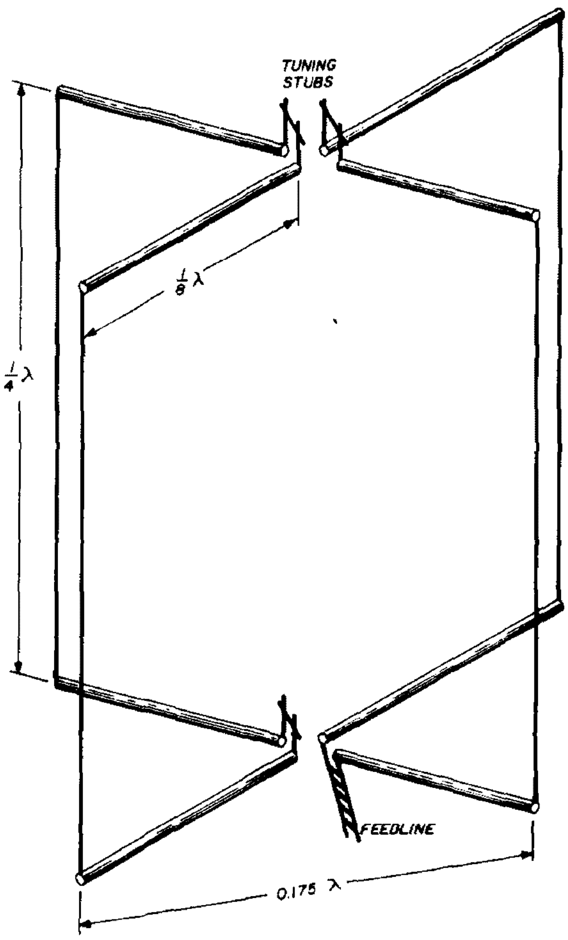



**fig. 1. G4ZU's quad design simplified construction and feedline problems.**

The German quad, a specialized high-gain multiband antenna designed by DJ4VM, appeared in the August, 1969, issue of *ham radio*. Editor

If the upper stubs are made from a ½-wavelength section of 300-ohm twin lead (at the desired 20-meter frequency), the antenna can also be used on 40 meters because the stub becomes ¼ wavelength at 7 MHz and the radiator becomes a half-wave element. The long stub can be run down inside the mast.

## swiss quad

Another radical departure from conventional cubical-quad design is HB9CV's Swiss quad shown in **fig. 2.** This antenna, like G4ZU's, has been more popular in Europe than here in the States, but it's an interesting approach that offers several constructional advantages.

The horizontal elements of the Swiss quad are made from aluminum tubing, the vertical elements are no. 14 to no. 10 stranded wire, and the antenna is fed with a gamma match and coaxial line. Complete dimensions for 10, 15 and 20 meters are given in **table 1.**

The vertical elements should be initially made slightly longer than that shown in **table 1** to facilitate tuning (this lowers the resonant frequency.) Couple a grid-dip meter to the antenna, and shorten the vertical wires to raise the resonant frequency to the desired point. A slight upward bend on the upper elements will keep the vertical wires tight.

After the quad is tuned to resonance check the swr. If it is too high adjust the gamma match accordingly. Be sure to lengthen or shorten both ends of the gamma match by exactly the same amount.

When building the Swiss quad it's extremely important that the mast be securely connected to the aluminum elements *and* well grounded. Some builders have found less tuning and swr problems by insulating the horizontal aluminum elements from the mast and connecting their exact centers with wire; this wire is then connected to the mast. The elements can be simply insulated by running the elements through short sections of flexible plastic water pipe.

## soviet quad

The unique three-band antenna shown in fig. 3 consists of two wire squares approximately 12 feet on a side, spaced 114 inches apart. The top and bottom of radiator while the other is a reflector; the reflector is fed through a 91-inch length of 300-ohm twinlead. According to the designer, UB5UG, the antenna has 8-dB gain on 14 MHz and 10-dB gain on 21 MHz.

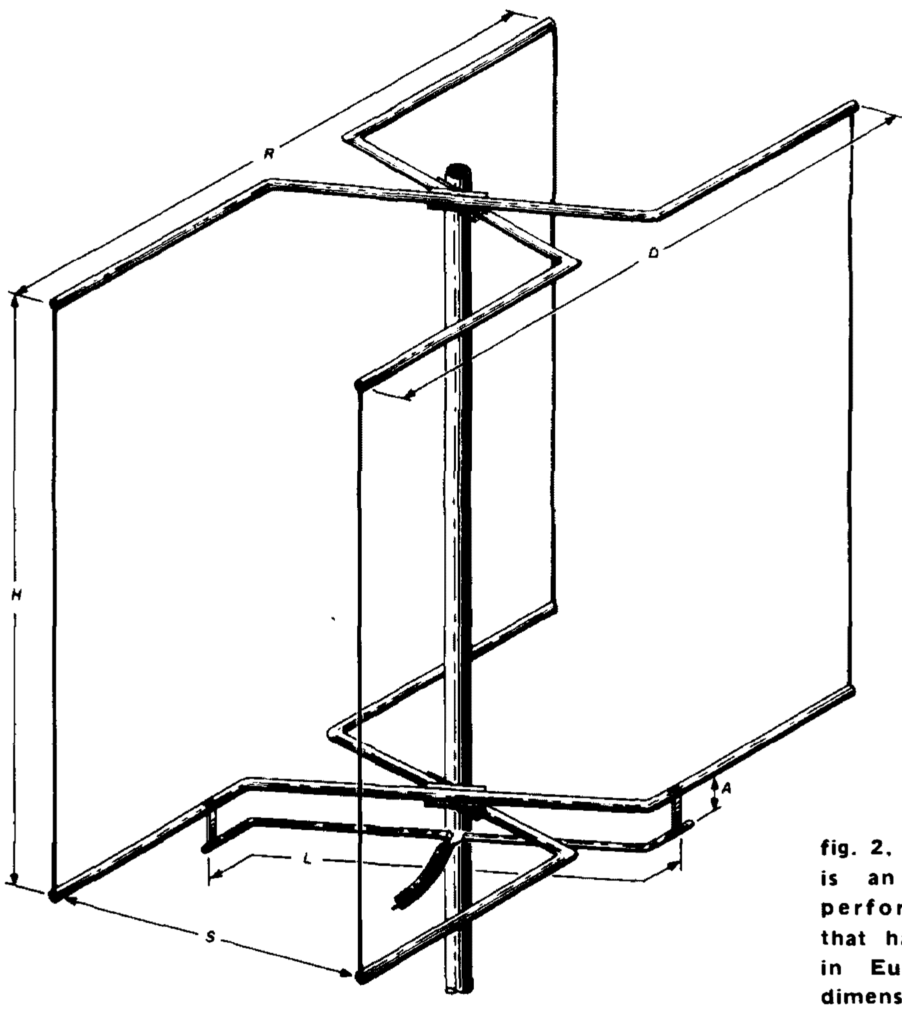



fig. 2. The Swiss quad is an unusual high-performance design that has been popular in Europe. Complete dimensions are given in table 1.

table 1. Dimensions for the Swiss quad (in inches).

| frequency (MHz) | reflector horizontal (R) | radiator horizonal (D) | vertical (H) | spacing (S) | gamma length (L) | gamma spacing (A) |
|---|---|---|---|---|---|---|
| 14.05 | 248 | 223½ | 236 | 84 | 188½ | 4¼ |
| 14.2 | 245 | 221 | 233 | 83 | 186½ | 4¼ |
| 21.05 | 165 | 149 | 157 | 56 | 126 | 2½ |
| 21.2 | 162½ | 148 | 155½ | 55½ | 124½ | 2½ |
| 28.05 | 124 | 112 | 118 | 42 | 94½ | 2 |
| 28.5 | 122 | 110 | 116 | 41½ | 93 | 2 |

each square is broken in the center and shunted with 300-ohm twinlead stubs that resonate the antenna on 10, 15 and 20 meters. One square is used as the

When fed with 50-ohm coax, the swr is less than 2.7:1 on all bands. Input resistance is 30 ohms on 14 MHz, 90 ohms on 21 MHz and 80 ohms on 28 MHz. For

lower swr performance, two equal-length pieces of 75-ohm coax can be used in parallel on twenty meters (equivalent impedance 37.5 ohms); for the upper two bands one piece of coax is disconnected at the transmitter. With this system the swr should be less than 1.5:1 on all bands.

Construction, as shown in **fig. 4** is somewhat different than that normally seen, although there is no reason a conventional quad arrangement could not be used. However, the UB5UG system doesn't use a long metal boom, and metal parts that might degrade performance are at the bottom out of the rf field.

Tuneup consists of adjusting the vertical 300-ohm stubs for resonance at the desired spot on the 20-meter band. Front-to-back ratio and forward gain are optimized by adjusting the length of the reflector feeder. (Be sure this feed line is twisted one-half turn.)

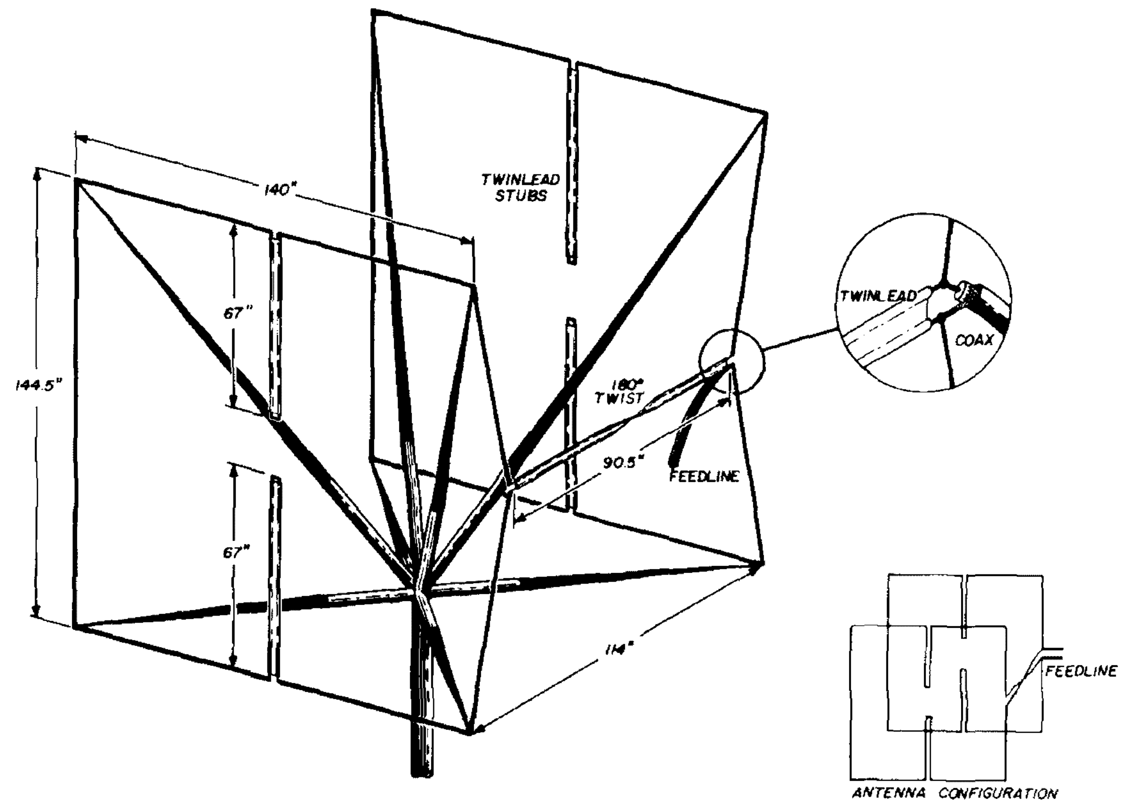



**fig. 3. UB5UG's simple quad provides three-band operation. Both reflector and director are the same size.**

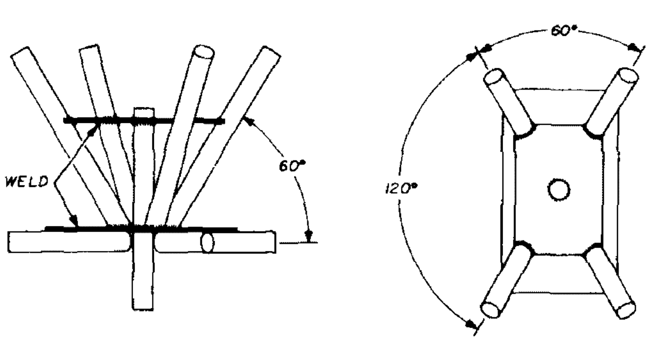



**fig. 4. Construction of the spider for the Soviet quad.**

## japanese quad

Cubical quads for 15 and 20 meters are pretty large, and some amateurs simply don't have the space to get them up into the air. The design in **fig. 5** by JA1OYY uses loading coils to reduce the size by 30%. Although the original design was for 15 meters, the idea is equally applicable to other bands; suggested coil values are lised in **table 2.**

A full-sized 21-MHz quad is 11 feet, 8 inches on a side; this reduced-size unit has sides that are 8 feet, 2 inches long. The

four loading coils are placed at low-current points so there is minimal effect on the radiation efficiency of the antenna. Spacing between radiator and director is 0.175 wavelength for maximum front-to-back ratio.

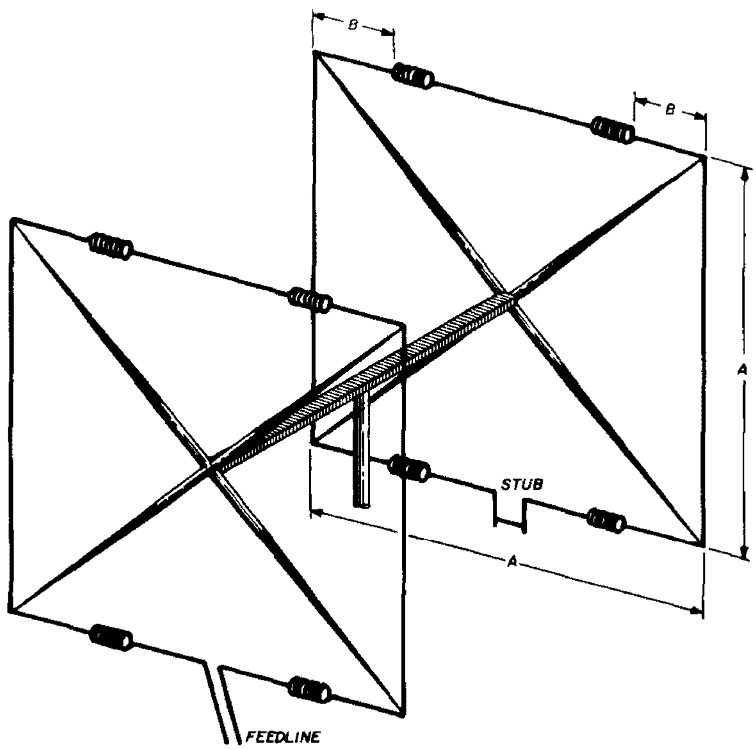



**fig. 5. Miniature quad design by JA1OYY uses loading coils in the horizontal elements. Complete dimensions are given in table 2.**

**table 2. Construction details of the Japanese quad (in inches).**

| frequency (MHz) | quad side (A) | spacing | coil location (B) | coil (μH) | winding info |
|---|---|---|---|---|---|
| 14 | 152 | 101 | 38 | 3.0 | 17½ turns no. 14, 1½" diameter, 5" long |
| 21 | 98 | 67 | 24½ | 2.0 | 11 turns no. 14, 1½" diameter, 3½" long |
| 28 | 77 | 50 | 19¼ | 1.5 | 9½ turns no. 14, 1½" diameter, 3½" long |

The loading coils consist of 11 turns no. 14 wire spaced to fill a form 1½ inches in diameter and 4 inches long. Commercial coil stock such as B&W or Airdux could also be used. The completed coils should be protected from the weather for long life and optimum performance; plastic refrigerator boxes are ideal for this purpose. Note that the coils on the bottom of the elements are equipped with alligator clips for tuneup. When the correct tap point has been found the clips can be replaced with soldered connections.

Although the quad built by JA1OYY uses angle-iron spiders and wood spreaders, fiberglass poles and cast aluminum spiders will last longer. With the 21-MHz dimensions shown in table 2 this miniature quad is resonant at 21.350 MHz (swr 1.1:1 with 50-ohm coaxial feedline); at 21.450 MHz the swr is 1.2:1 at 21.200 MHz the swr is 1.5:1.

**ham radio**

# multiband dipoles for portable use

Low cost, simplicity, and efficiency are the keynotes of these antenna designs

William I. Orr, W6SAI, Eimac Division of Varian, San Carlos, California 94070

**The multiband dipole** is an attractive antenna for portable or Field Day operation. It's light, easy to carry and erect, and very inexpensive to build. Best of all, it's noncritical in adjustment and provides a low swr for commonly used 50-ohm coaxial transmission lines.

Simple multiband dipole antennas can take one of two forms. The most easily constructed two-band antenna consists merely of two dipoles tied in parallel at the feed point. A more sophisticated antenna uses parallel-tuned traps in series with the dipole element (fig. 2).

It was desired to build a few antennas for the ARRL Field Day contest, but a search of the literature failed to provide any dimensional data or information on multiband dipole systems for the higher-frequency bands. Accordingly, several multiband dipole antennas were constructed and tested during the spring of 1968 with the hope of obtaining a suitable design before the contest began. This article covers three designs that evolved from these tests. Perhaps one of them will suit your Field Day requirements.

## parallel dipoles

The parallel dipole shown in fig. 1 is simpler than the trap dipole, but its use is limited to *harmonically related* bands. A parallel dipole may be built for 20 and 10 meters, because 10 meters is the second harmonic of 20 meters. However, trouble occurs when an attempt is made to use two dipoles not harmonically related. For example, a parallel dipole for 20 and 15 meters or for 15 and 10 meters isn't practical. These bands aren't harmonically related, and the dipoles have so much mutual reactance it's impossible to establish resonance without an auxiliary tuning unit. This defeats the whole purpose of the undertaking; namely, that of constructing a *simple* multiband antenna system.

For harmonically related bands, however, the lower-frequency dipole presents a high resistive impedance to the higher-frequency dipole and doesn't affect its operation to any noticeable extent. In the case of *harmonically unrelated* bands, such as 20 and 15, or 15 and 10 meters, the lower-frequency dipole is highly reactive at the higher operating frequency and introduces an intolerable degree of detuning to the higher-frequency antenna. The higher-frequency antenna doesn't influence the lower-frequency antenna to any great degree, but the lower frequency antenna—unless it presents a

resistive load to the higher frequency antenna—introduces sufficient reactance at the feedpoint of the combined system to "upset the applecart."

Paralleled, harmonically related antennas detune each other to a small extent. The low-frequency dipole appears to resonate at a lower frequency and the higher frequency dipole at a higher frequency than normal. This can be compensated by trimming the dipoles to their operating frequencies.

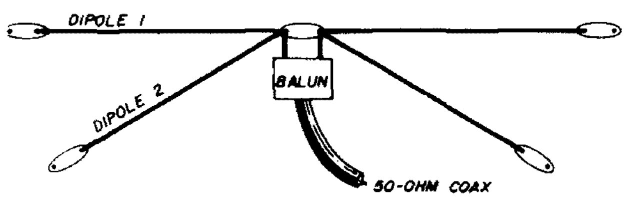



**fig. 1. Parallel dipole arrangement. Two dipoles, resonant at harmonically related bands, are connected in parallel; interaction is negligible because of impedance relationships.**

## trap dipoles

The trap antenna operates on the principle of parallel-tuned circuits placed at critical points in the element. The tuned circuits or traps, electrically connect or disconnect the outer sections of the element as the antenna's excitation frequency is changed.

At the lowest operating frequency, the tuned traps have a minimum effect on the antenna, which is resonant at a frequency determined by its electrical length. The electrical length is influenced slightly by the loading effect contributed by the inductive reactance of the traps. At some higher frequency, the outer set of traps is parallel resonant. This places a high impedance between the center portion of the element and the ends beyond the traps. Thus, the element resonates at a frequency higher than that determined by the over-all physical length of the element. As the operating frequency is increased, the inner set of traps becomes resonant, effectively disconnecting a larger portion of the element from the center driven section. The length of the center section is therefore resonant at the highest operation frequency. The center section, plus two adjacent inner sections, are resonant at the intermediate operation frequency, and the complete element is resonant at the lowest operation frequency.

At the lower frequencies, the traps in the active portion of the element contribute a reactive load that makes the over-all element length somewhat shorter than that determined by formula. As a result, the lengths of the multiband trapped dipole sections must be determined by cut-and-try, since the loading contributed by the traps is difficult to determine.

The efficiency of such a system is generally established by trap- and element-tuning accuracy and by the Q of the traps. An accurately tuned multiband dipole using high-Q traps compares favorably with full-size, separate dipoles as far as efficiency and bandwidth are concerned.

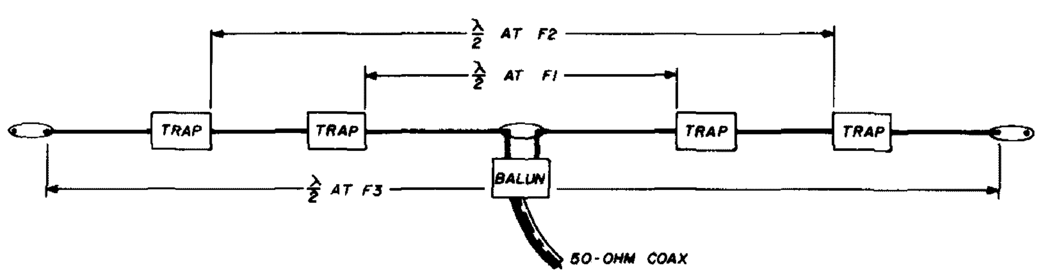



**fig. 2. Trap dipole for nonharmonically related bands. Traps are parallel-resonant circuits that provide high impedance at resonant frequency and low impedance at frequencies far from resonance.**

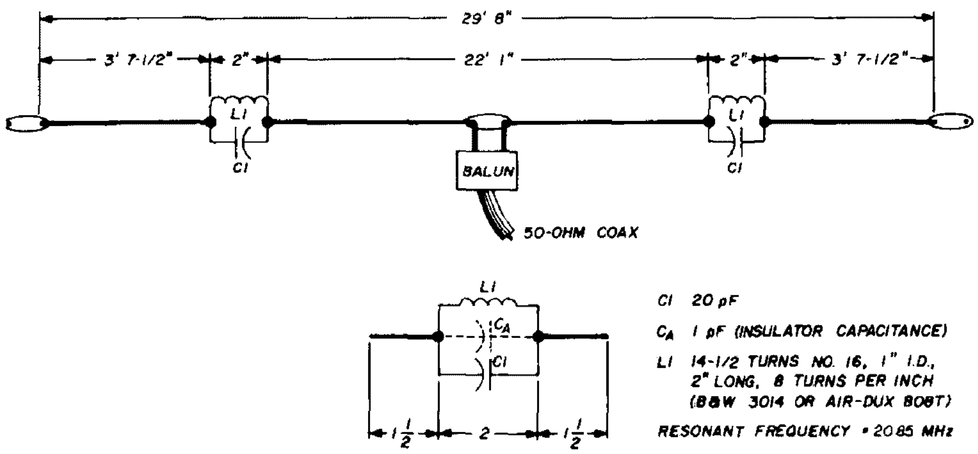
## 20-15 meter trap dipoles

These bands aren't harmonically related, so a parallel dipole isn't practical, and a trap dipole must be used **(fig. 3)**. The dimensions are given in **fig. 3A.** The 15-meter portion is resonant at about 21.15 MHz. It's length is very close to the figure obtained from the formula given in the handbooks.

The 20-meter length is shortened by the loading effect of the 15-meter trap to 29 feet, 8 inches. The traps are about 2 inches long. This length must be included in the over-all measurement, which makes the tip sections each 3 feet, 7½ inches long.

The dipole is fed from a balun.[1] This is necessary if meaningful swr measurements are to be obtained. The feedline may be a random length of RG-58A/U for power up to 500 watts PEP, or RG-8/U for power up to 1000 watts PEP. The smaller cable is recommended for Field-Day use, as it's easier to handle than the RG-8/U.

The completed 20-15 meter trap antenna produces the swr curves shown in the chart of **fig. 3C.** At the design frequency of 14.15 MHz, the 20-meter dipole exhibits an swr of about 1.25 at the end of a half wavelength of transmission line. At the limits of the 20-meter band, the swr rises to less than 1.6. On 15 meters, at the design frequency of 21.15 MHz, the antenna exhibits an swr of about 1.09, rising to less than 1.9 the high-frequency end of the band and to 1.4 at the low-frequency end. These swr figures are well within the limits normally established



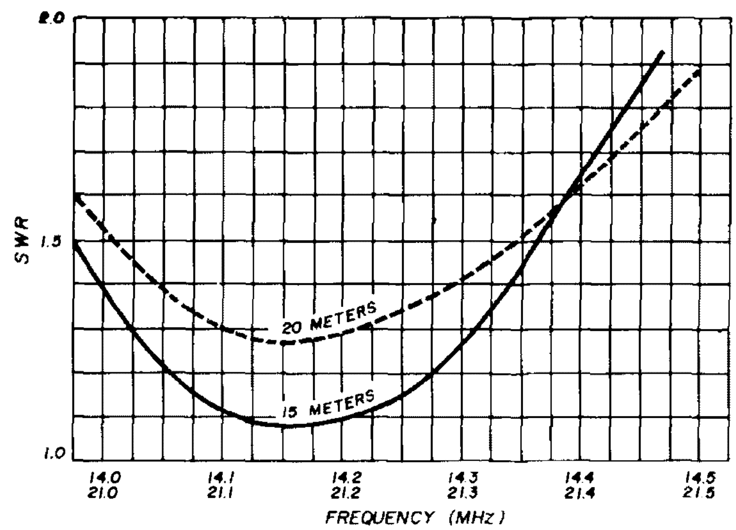




**fig. 3. Design data for a dual-band 20-15 meter trap dipole including trap construction details and swr response.**

for use with most transmitting equipment.

## 20-10 meter parallel dipole

A practical 20-10 meter parallel dipole is shown in **fig. 4.** Two separate dipoles are connected in parallel at the feed point, with their ends separated about five feet. The 20-meter dipole dimensions are normal but the 10-meter dipole is lengthened slightly to re-establish resonance, which is affected by the addition of the 20-meter section. Formulas for computing correct dipole lengths for other harmonically related bands (40 and 20 meters, for example) are given. Dipole spacing at the ends isn't critical; however, too-close spacing will detune the antennas. A separation of five feet is satisfactory for the 20-10 meter antenna system; for a 40-20 meter antenna, ten feet is satisfactory.

Again, a simple balun is placed at the antenna, as suggested for the previous antenna. Typical swr curves for the 20- and 10-meter bands are shown in **fig. 4B.** The swr remains below 1.6 across the 20-meter band and remains below 1.6 across nearly a megahertz of the 10-meter band. For the high end of 10 meters, the 10-meter dipole should be shortened about three inches at each end.

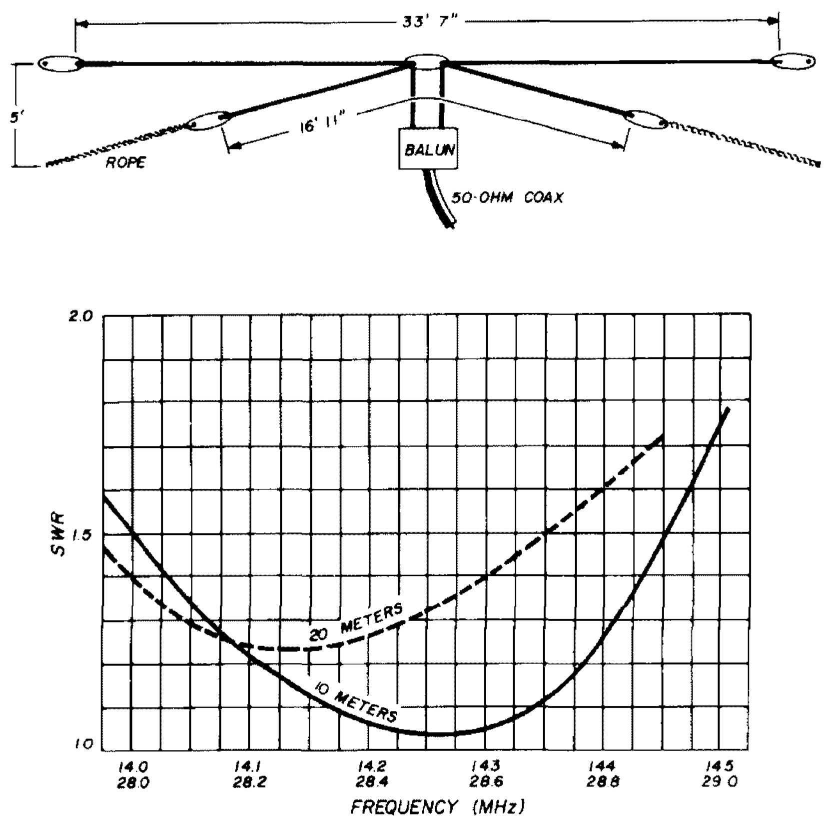



**fig. 4. 20-10 meter parallel dipole design and swr as a function of frequency. No traps are needed in this inexpensive system, which can be adapted for other harmonically related bands. The length of the lower-frequency dipole (in feet) is equal to 478/f where f is a frequency in MHz; length of the higher-frequency dipole is 485/f.**

## a 20-15-10 meter trap dipole

The trap technique may be extended to three bands. Two sets of traps are required, one for 10 meters and one for 15 meters. Construction information for a triband trap dipole is given in **fig. 5.** While the 10-meter dimensions are normal, both antenna lengths for the 15- and 20-meter sections are shortened by trap loading. Over-all antenna length is only 26 feet, 2 inches. No matter. The antenna performs remarkably well on all three

bands, and comparisons with full-size dipoles indicated little or no difference in signal strength between antennas, either on transmit or receive.

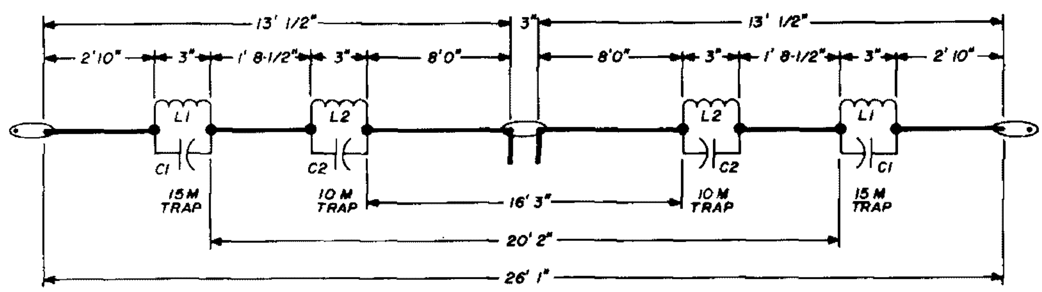
Two sets of traps decrease the operaing bandwidth somewhat, as seen from the swr curves of **fig. 5B**. However, the over-all bandwidths still permit an swr of 2.15 or less across the 20-meter band, with a minimum swr of 1.3 at the resonant frequency of 14.3 MHz. The 15-meter antenna meter swr is higher. It's close to 1.6 at the resonant frequency of 21.22 MHz, rising to about 2.4 at the lower band edge and to 2.2 at the high edge. The 10-meter swr curve is quite broad. Although not quite as good as that of a single 10-meter dipole, it's about the same as that of the parallel dipole shown in **fig. 4**. Nevertheless, the swr remains below 1.6 from the low edge of the phone band to about 28.85 MHz, and is below 2.0 to 29.0 MHz.

As in the case of the other antennas, a balun feed and 50-ohm transmission line are used.

## trap resonant frequency

Trap adjustment is easy if the proper procedure is used. The only instruments required are a grid-dip oscillator and a calibrated receiver. But you must know *where* to tune the traps for optimum performance.

Tradition says the traps should be tuned to the low edge of the band, or just outside the low edge. Measurements made on various trap beam antennas showed that the traps were indeed tuned in this manner. Discussions with antenna manufacturers on trap tuning have led to the conclusion that traps are adjusted in this manner merely because the competition does it, and none really knows the reason



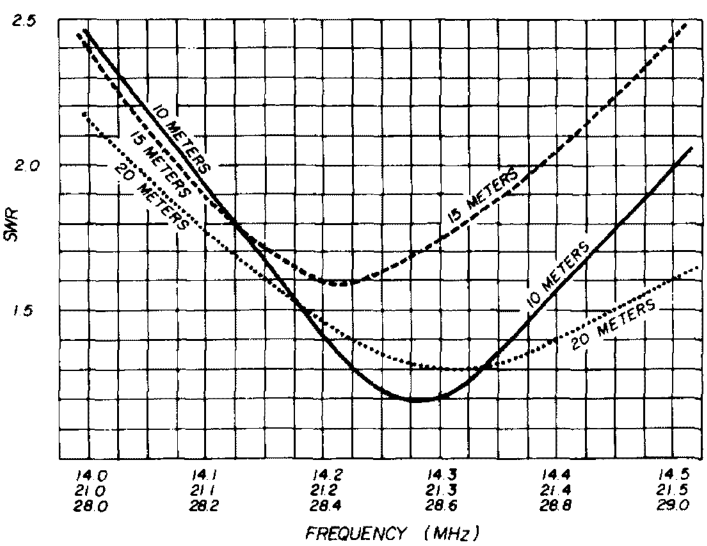
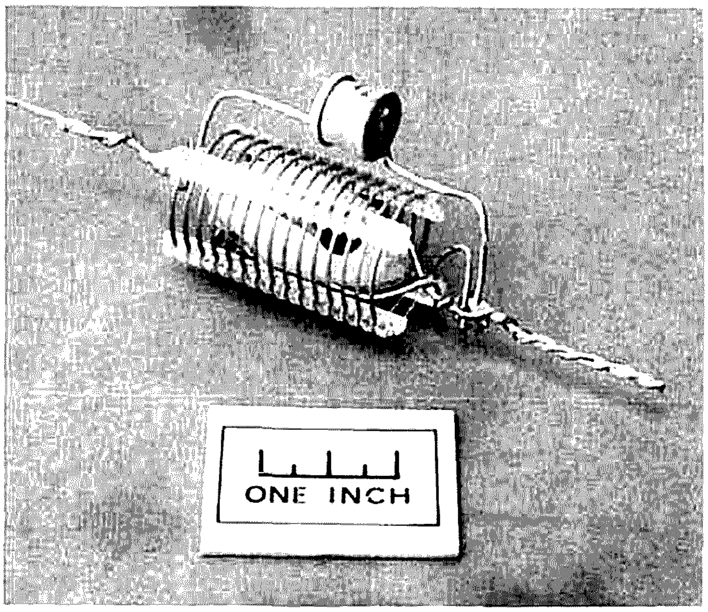




**fig. 5. Trap dipole for 20-15-10 meters and swr response. The 15-meter trap is the same as in fig. 3; for 10 meters, use 20 pF in parallel with 8-3/4 turns number 16 wire, 1 inch ID x 1-1/8 inch long, 8 turns per inch (or use Air-Dux 808T). Resonate to 27.8 MHz.**

for trap adjustment–if they do, they're not saying! I saw no valid reason for this particular choice of trap resonant frequency, so I made some tests.

The tests revealed (1) there's no magic frequency to which the traps should be tuned, and (2) if the trap is tuned to resonance with the band in which it is supposed to resemble a switch, all will work as expected.

However, I noticed on repeated tests that, when the trap was tuned *lower* in frequency than the expected range of




**Typical 15-meter trap. Lead length affects resonant frequency slightly, and trap should be measured with connecting leads in place.**

operation, the antenna bandwidth on that band was improved. In other words, tuning the trap to some frequency within the band restricted antenna bandwidth. To obtain the best bandwidth, it proved best to tune the traps somewhat *lower* than the lowest expected operating frequency.

The exact frequency to which the trap was tuned didn't seem too critical as long as both traps (for the same band) were tuned to the same frequency. Tests using tunable traps indicated good results with the 10-meter traps tuned to 27.8 MHz, the 15-meter traps tuned to 20.85 MHz, and (for 40-20 meter trap antennas) the 20-meter traps tuned to 13.9 MHz. As long as the traps were resonant *outside* the low-frequency end of the band by fifty to several hundred kilohertz, antenna action remained unchanged for all practical purposes. When the trap resonant frequency approached the operating frequency, or was higher than the operating frequency, antenna bandwidth became progressively more restricted, as shown by the measured standing wave ratio.

This seemed to be the "secret" of proper trap adjustment. Just *why* this is so I'll leave as an exercise for the reader.

## trap construction

Construction of a typical trap is suggested in the photo. The trap can be built around a strain insulator, which will remove the antenna tension from the coil and capacitor. The strain insulator has measurable capacitance (about 1 pF) so it must be considered as part of the trap circuit. The coil may be placed around the insulator, with the capacitor connected around the coil.

Trap length enters into the over-all length of the antenna. A standard length of two inches is used for the traps shown here. Generally speaking, the L/C ratio of the trap is not too important; traps have been built around high-voltage ceramic capacitors of 20 or 25 pF. For power output of 500 watts PEP, the Centralab 853A-20Z ceramic capacitor is ideal. It has a capacitance of 20 pF and a voltage rating of 3 kV. Small, commercially built air wound inductors may be used for the trap coils, such as the Barker & Williamson or Air-Dux coils.

To attach the traps easily to the antenna wires, it's a good idea to add 1½-inch leads to each end of the trap. The leads may then be twisted around the antenna wires and soldered.

## assembling the traps

Attach about a foot of wire to the strain insulator, then loop the wire through the insulator and back upon itself. Cut the wire to length and solder the joint. The coil may be placed over the

insulator or beside it. Solder the coil leads to those of the insulator, then solder the capacitor in parallel.

### trap tuning

Place the trap in a clear spot away from metal objects. Loosely couple it to a grid-dip oscillator. Check the trap resonant frequency against a calibrated receiver. As mentioned previously, the exact resonant frequency isn't important as long as it's outside the low-frequency end of the band. Both traps of the set should have resonant frequencies within about 50 kHz of each other.

### weatherproofing

Moistureproofing seems to be more of a problem than building the traps. If it's guaranteed not to rain on Field Day, the traps may be used without a protective shield. An emergency rain shield may be made by wrapping the traps in kitchen plastic and stapling the plastic cover around the trap. A better shield may be made from a polyethylene "squeeze bottle," obtainable at a drug store. The best (and most time-consuming) trap shield may be made of plastic tubing with end pieces cemented into a waterproof cylinder around the trap.

Regardless of the weatherproofing method, it's imperative that the traps be protected from water, otherwise the traps will become detuned when wet and may, in fact, flash over and be destroyed.

### antenna adjustment

The antenna dimensions and trap data guarantee that an antenna constructed in this fashion will *work.* A check on operation may be made by running an swr curve across the amateur bands for which the antenna is designed. The resulting curves should resemble those in this article. Note, however, that swr curves obtained *without* a balun between the antenna and the feedline are meaningless and are not representative of antenna performance.

If the swr on one or more bands indicates the antenna is resonant at a frequency other than that desired, antenna resonance may be re-established by altering the length of the antenna section in question. Adjustments shouldn't be made to the traps after once set, as too many variables will certainly lead to improper antenna adjustment.

Adjustment should be made to the highest-frequency antenna first, followed by the next lower-frequency segment. Once the 10-meter sections are of the proper length, the 15-meter sections may be adjusted, then the 20-meter section. It must be remembered that adjustments made to *one* antenna section affect the adjustments of the *lower*-frequency sections, since the higher-frequency antennas form a part of the lower-frequency ones.

Don't deliberately look for trouble, because the data and measurements given for these antennas have been tried in various installations with complete success. It can be seen, futhermore, that trap antennas can be designed for two or more frequency bands, regardless of their harmonic relationship, in the fashion outlined in this article.

### power limitations

It's possible to destroy an antenna trap, as some "California kilowatt" users have discovered to their dismay. The limiting factors are capacitor flashover and coil heating. The trap construction described in this article is satisfactory for power up to 500 watts PEP in the antenna; but for the maximum amateur power rating, it's recommended that more robust components be used. For high-power construction, two Centralab type 858, 50-pF, 5-kV capacitors may be connected in series to provide a 10-kV rating. The coil could be self supporting, wound with tv aluminum ground wire or number 10 copper wire. Trap size will increase, requiring slight adjustments to the antenna sections.

**reference**

1. William I. Orr, "Broadband Antenna Baluns," *ham radio,* June, 1968, p. 6.

**ham radio**

# improved triangular shaped beam antenna

Norman B. Watson, W6DL

Presented in this article is a beam antenna covering 7 through 28 MHz using triangular-shaped elements. It is simpler than similar antennas I have developed previously.[1, 2] Instead of 14 elements, only 6 are used. Little if any sacrifice in gain occurs on 21 MHz, and there's an increase in gain on 28 MHz compared with the original design.

## electrical description

All elements except the 7- and 21-MHz driven element and director are closed loops (**fig. 1**). The 7-MHz beam consists of a driven element and director spaced 1/8 wavelength. The wire elements are folded in a triangular shape with sides of equal length. The bottom leg of the triangle is parallel to the earth. The driven elements are fed at the center of this horizontal side. Nominal radiation resistance of the 7-MHz antenna is 15 ohms.

The 21-MHz antenna uses the same two elements as the 7-MHz antenna. Since its operating frequency is three times the fundamental, the 21-MHz antenna's electrical length is 1.5 wavelengths. Element spacing is 3/8 wavelength. Its radiation resistance is 215 ohms.

The 14-MHz antenna is a four-element beam. Reflector, driven element, and two directors are closed triangular loops one wavelength in perimeter. The elements are spaced about 1/8 wavelength. The driven element is fed at the center of the triangle horizontal leg in the same way as the 7/21-MHz driven element.

Physically, the 28-MHz antenna doubles as the 14-MHz antenna. Electrically, however, it is two wavelengths in perimeter on 28 MHz. Element spacing is ¼ wavelength on this band. Radiation resistance on 28 MHz is 175 ohms; on 14 MHz it is 65 ohms.

## directivity and gain

The beam's directivity on the four bands is given in **table 1**. Directivity, as considered here, is a measure of **beamwidth**. That is, the directivity is the power concentrated through a portion of the space around the antenna, as opposed to that from an isotropic source.

Directivity is a useful comparative parameter. It's merely necessary to plot a receiving antenna pattern using a signal source at any desired distance from the antenna, then calculate the directivity using the half-power beamwidth. The numbers in **table 1** are based on the

assumption that antenna beamwidths are the same in the vertical and horizontal planes.

## wire lengths

The wire lengths in **fig. 1** were determined by cut and try. The resonant frequency of the driven elements was measured at an operating height of 60 feet above ground, using an ac frequency-compensated bridge. Each antenna element is a series-resonant circuit, whose resonant frequency is a function of

| FREQUENCY BAND (MHz) | NUMBER OF ELEMENTS | ELEMENT SPACING (APPROX λ) | DIRECTIVITY (SEE TEXT) * | RADIATION RESISTANCE (Ω) |
|---|---|---|---|---|
| 7 | 2 | 1/8 | 4 | 15 |
| 14 | 4 | 1/8 | 10 | 65 |
| 21 | 2 | 3/8 | 7 | 215 |
| 28 | 4 | 1/4 | 16 | 175 |

*AS A BASIS FOR COMPARISON, THE DIRECTIVITY OF A HALF-WAVELENGTH ANTENNA IS 1.64

**table 1. Beam antenna characteristics**

wire lengths of the driven elements. An over-all change of 13 inches will move the resonant frequency about 0.1 MHz and 0.22 MHz in the 7- and 14-MHz bands respectively. The other element lengths

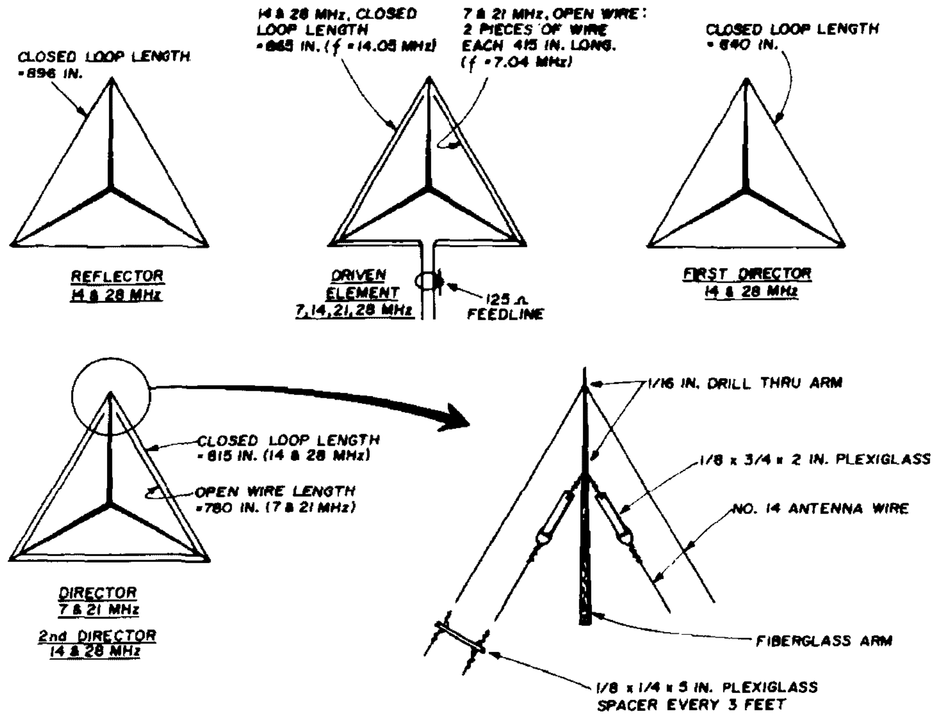



**fig. 1. Layout detail of the triangular elements.**

physical wire size and length. Capacitance is determined by the physical distance between the element and the ground, tower, houses, trees, antenna boom, and other antenna elements.

For these reasons, wire-element lengths will be affected by your antenna site and by antenna height. The antenna should have the wire lengths given in **fig. 1**. The resonant frequency of the driven elements should be measured at the operating height. (A measurement method is described below.)

If the operating frequency is unsatisfactory, it may be altered by changing the

can remain unchanged with little effect on performance.

## feed line type and tuning

A single 125-ohm shielded transmission line feeds both driven elements, which are tied together at a common feed point. The feed line is electrically one wavelength long at the operating frequency in the 7-MHz band, which makes it 2 wavelengths on 14 MHz; 3 wavelengths on 21 MHz; and 4 wavelengths on 28 MHz. A feed line that is a multiple of ½ wavelength reflects the load resistance to the input end of the line. Therefore,

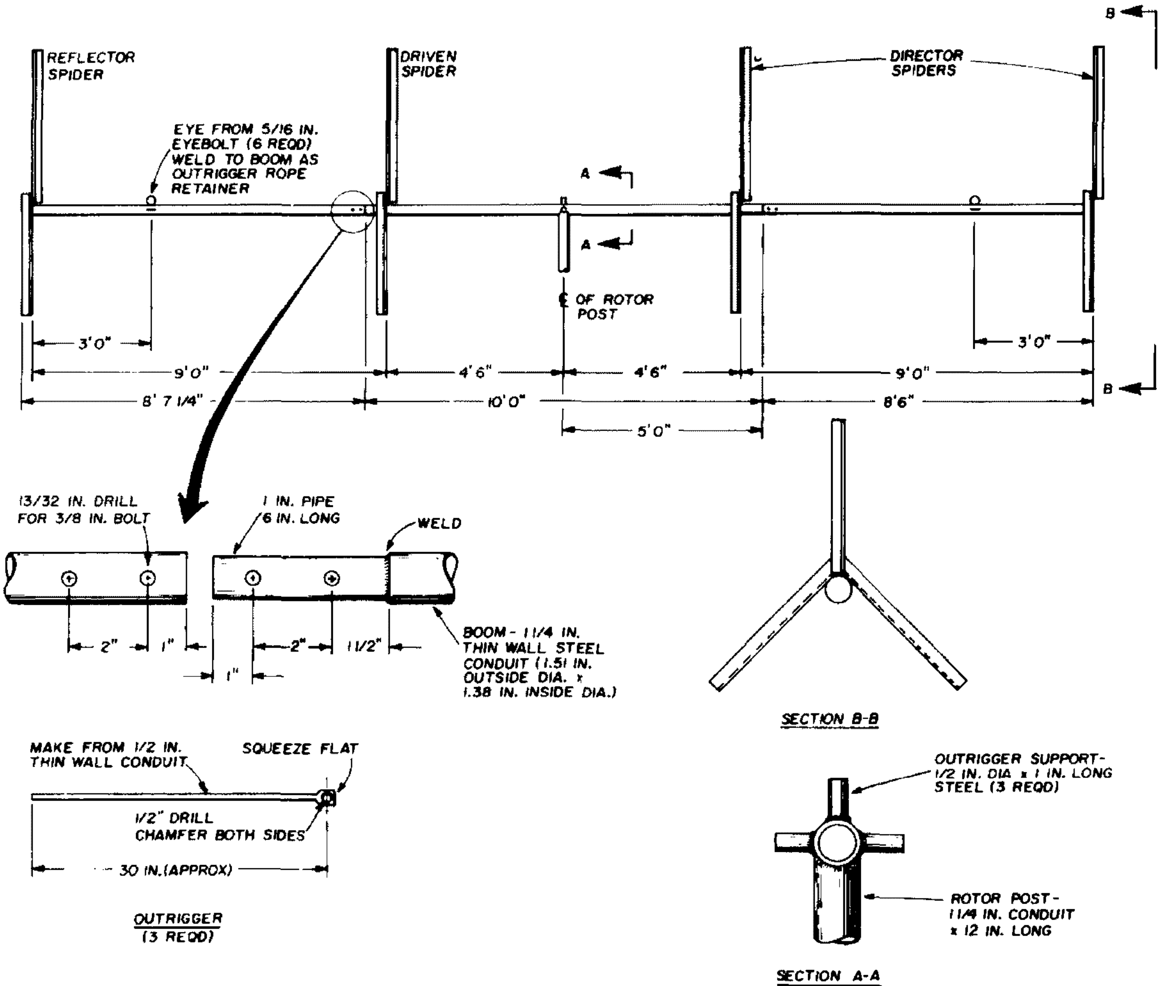



**fig. 2. Construction of the boom.**

the even-wavelength-multiple line makes it possible to operate the antenna on four bands with a single feed line, providing an antenna tuner is used to match the transmitter to the resistance appearing at the input end of the line. This resistance will be the antenna's radiation resistance, providing the antenna is tuned (cut) for resonance at the frequency at which the feed line is a multiple of ½ wavelength.

Here's a procedure for tuning the antenna-feed line system:

**1.** Decide on the frequency at which optimum efficient is desired. This will be influenced by whether cw or phone operation is contemplated for the greater portion of time.

**2.** Calculate the approximate electrical length of feed line needed for operation at the desired frequency. Consider the line's velocity factor and the distance from transmitter to antenna feed point. (A discussion of velocity factor is contained in the ARRL Handbook.) Add a couple of feet to the calculated length as a safety factor, and cut the line.

**3.** Short circuit one end of the line, and connect the antenna bridge to the other end. Determine the frequency at which a null occurs with the bridge set to zero. Cut the line, a little at a time, until the bridge nulls at the desired operating frequency. The greater the number of half wavelengths in the line, the less will be the frequency change per inch of length removed.

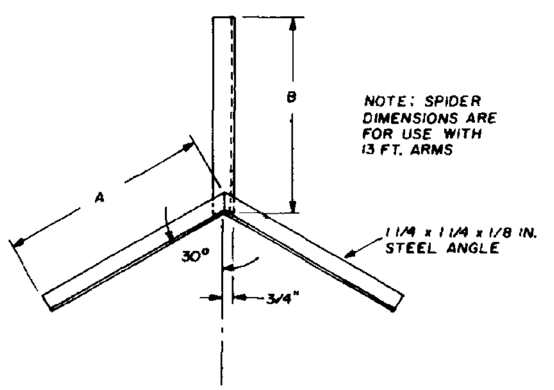
**4.** Run the line from the operating position up the tower and out the antenna boom to the driven element spider. From the driven element spider, the line can hang free to where it's attached to the driven elements. Connect the feed line to **both** driven

elements, not to just one. (Interaction exists between all elements on this antenna as with any beam.)

5. Connect the bridge to the operating-position end of the transmission line. Raise the tower and antenna to the operating height. Find the bridge null that indicates the antenna resonant frequency. Alter the antenna driven element length to bring its resonant frequency to the desired operating frequency. Antenna radiation resistance will be reflected at the bridge-end of the line when antenna resonant frequency is the same as that at which the transmission line is a multiple of an electrical half wavelength.

6. Connect the bridge to the antenna tuner output. Set the bridge to the antenna radiation resistance of one of the bands. Tune the antenna tuner to the operating frequency for the band,

**fig. 3. Spider.**




| ELEMENT | A (INCHES) | B (INCHES) |
|---|---|---|
| REFLECTOR | 26 1/2 | 27 |
| DRIVEN | 20 | 21 |
| DIRECTORS | 16 | 17 |

as indicated by a null of the bridge. Record the tuner dial settings for use when changing bands. Repeat the procedures for each band.

## construction

My original 4-element beam was in use for three years. The experience gained over this period has resulted in simpler construction of the beam described here. Except for a few pieces of scrap Plexiglas, all materials can be found in the Sears mail-order catalogue. Simple hand tools are required, and some welding is necessary. The advantage of welded construction has been proven with my original large beam. No failures have occurred in three years of storms howling in from the Pacific Ocean.

The 27-foot boom is constructed from three 10-foot lengths of 1¼-inch thin-wall steel conduit (**fig. 2**). The center section is 10 feet long. Cut off the two end sections with a hacksaw or tubing cutter at 8 feet 6 inches and 8 feet 7 inches respectively. Weld a 6-inch length of 1-inch pipe to each end of the center section to serve as a joint. Assemble the sections using 3/8-inch stainless-steel bolts.

The four spiders are constructed as shown in fig. **3.** Clamp the two "A" legs to a piece of scrap wood before welding: make sure the angles are correct. Then clamp the "B" leg to the assembly and weld. Next, weld the spiders to the boom. The "V" joint of the spider sets on top of the boom (section B-B, **fig. 2**). Three different spider lengths, fig. **3**, are used to save material and weight.

Weld a 12-inch length of 1¼-inch conduit to the center boom section to support the beam on the rotor. Don't increase the length of this center section as a means of increasing the beam height above ground. The 12-inch long rotor section is strong enough to withstand high winds with ice. It can be made longer for attachment to a rotor located down in the tower, providing an upper bearing is located within 6 inches of the boom.

Cut three one-inch-long pieces from a ½-inch bolt, and weld them to the top and sides of the boom to support the outriggers as shown in **fig. 2.** Cut three pieces 31 inches long from a 10-foot length of ½-inch thin-wall conduit for outriggers. Squeeze the end of the outrigger together in a vise and drill a ½-inch hole through the flat section. The 3/8-inch outrigger rope will be threaded through these holes when the beam is

assembled. Cut the eyes off of six 5/16-inch eyebolts, and weld the eyes to the boom as shown in **fig. 2.** The outrigger ropes will be fastened to these eyes. After welding and drilling are completed on the boom and outriggers, they should be hot-dip galvanized. Galvanizing is nominal in cost. The entire job won't cost more than four or five dollars, and galvanized parts will last for many years.

All that remains to be made now are the antenna-to-feedline connector plate, **fig. 4**, and the Plexiglass insulators and spreaders, as shown in **fig. 1.** Cut a notch in each end of the spreader to fit over the

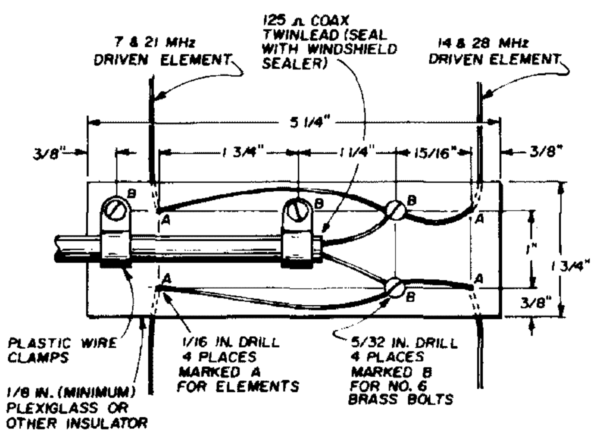



**fig. 4. Antenna-to-feedline connector plate.**

antenna wire, and drill a hole in each end for wiring the spreader to the antenna.

Automobile hose clamps are used to fasten the fiberglass arms to the spiders. Twenty-four clamps are required. Solid braid polyethylene rope is used for the outriggers. It won't stretch, so no turnbuckles are required. Polyethylene rope is used extensively for boats and can be obtained from the Sears mail order catalogue. Three pieces, each 23 feet 6 inches long by 3/8-inch diameter are required. After cutting, fuse the end over a gas flame.

Thirteen-foot-long fiberglass arms are used for wire spreaders. Twelve are required. Obtain the most rigid arms you can find. They should be at least ½-inch diameter at the small end and 1-3/8-inch diameter at the large end. The arms I used are marketed by Kirk Electronics and manufactured by U. S. Fiberglass.

## assembly

After the boom and outriggers are galvanized, assemble the boom section on the ground. Clean any excess zinc from mating sections and bolt holes.

If you have a tilting tower, as I do, assembly is very straightforward. Assemble the fiberglass arms to the two center section spiders. With one set of arms resting flat on the ground, string the antenna wires and make the solder joints. Tilt the tower down and attach the boom to the rotor. Raise the center section out of the way. Assemble the fiberglass arms to the two end section spiders, with the arms flat on the ground and the boom section joint pointing upward. String the wires on the end element supports. Tilt the tower down, and bolt an end section to the center section. Tie the three outrigger ropes to the boom eyes. The length of outrigger rope given allows a foot at each end for tying to the eyes.

Raise the antenna, rotate the boom 180°, and tilt the tower down again. Thread the outrigger ropes through the outriggers and insert the outrigger tubes over the ½-inch-diameter retainers. Raise the tower so the second end section can be attached to the boom. Tie the outrigger ropes to the second set of eyes. The side outrigger ropes need only be snugly tight with the boom straight. The top outrigger may have to be readjusted after the tower has been raised to ascertain if the boom is level.

If you have a rigid or crank-up tower, you're on your own for an assembly method.

If you'd like to start building a less complex antenna than this beam, try only the driven frame. Feed it in the same manner as described above. It will function as an excellent four band antenna.

I'll be glad to help serious builders with further technical information.

### references

1. Norm Watson, W6DL, "Triangular Loop Antenna," *QST*, April, 1968.

2. Norm Watson, W6DL, "Triangular Loop Beams," *73*, May, 1968.

**ham radio**

# 80 meters vertical beam antenna

George Cousins, VE1TG

Just "getting Out" on 80 meters calls for nothing too elaborate in the way of antennas. But to really get out, to the tune of working a world-wide DX as a matter of course, calls for something a bit special. The antenna described here is the result of a fair amount of experimentation and construction on the part of Bill Shearman, VE1AX, whose DX countries total on 80-meter ssb is well over the magical 100 mark.

To say it's a two-element beam conjures up ghastly thoughts of huge towers and monstrous drooping elements, but this beam is vertical, which makes all the difference in the world. For one thing, it's unobtrusive. It sits right in the middle of a subdivision and is not particularly noticeable. Being a quarter-wave vertical, the elements are only about 60 feet high. Although this may seem a lot, by using small-diameter masts the problems of height and weight become quite reasonable.

## elements

The antenna at VE1AX consists of two identical 48-foot aluminum masts topped with a 12-foot fiberglass whip of the type used for marine transmitters. Any suitable combination of aluminum or steel tubing, tv masts, etc. could be used; but both verticals must be identical. The masts must be insulated from ground, which Bill did by using commercially built tripod mounts. But here again, traditional ham ingenuity should come to the rescue with such alternates as wooden posts, power line standoff insulators, or even Coke bottles, as I remember seeing in another article.

## broadbanding

To make the antenna broadband, a three-wire cage encloses each mast. Fig. 1 and the photo show that these wires extend out from the mast about eight feet, and are connected to the mast at a point about fifteen feet from the top. (The exact point isn't critical.) Then they extend downward to the base of the mast. The upper arms of the cage connect to strain insulators, which are connected to the guys. The lower ends of the wires are also connected to strain insulators, which are in turn tied to the base mounting. This arrangement keeps the cage neat and taut. All the cage wires are connected electrically to the base of the vertical element and fed in parallel with it.

**One of the vertical beam elements. Loops near insulators are bridge wires connecting horizontal and vertical cage wires.**

A small loading coil is connected in series with the coaxial cable feeding the base of each vertical, and a movable tap is used to adjust the element for optimum performance.

## ground system

Like all vertical antennas, the secret of this one is the ground system. The old axiom of, "the more, the better," couldn't be more true. Bill's philosophy is to lay out every bit of spare wire he can

get his hands on, and the ground under and around the array is a maze of everything from hookup wire to braided ground strapping. At this writing, well over 1000 feet of radials have been laid out, but there will likely be more by the time you read this! Each little bit helps put the signal where it will do the most good – low down, and a long way out.

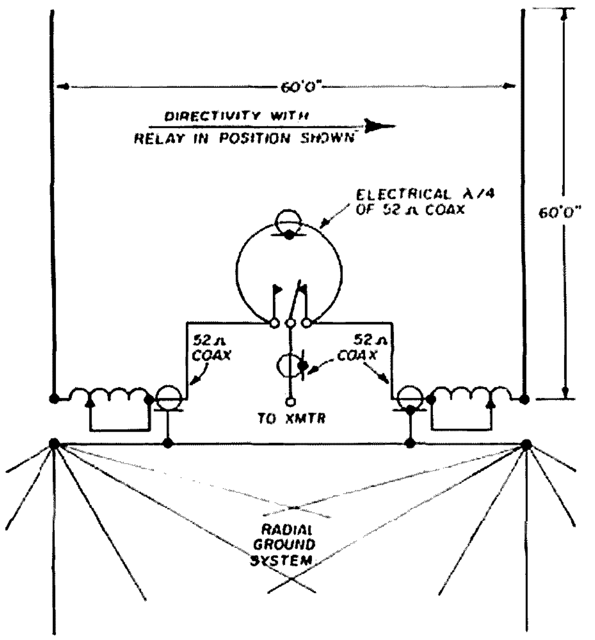



**fig. 1. General construction of the array. Loading coils are made from commercial coil stock or home wound, approximately 35 turns, 2½ inches in diameter, close-spaced.**

## orientation and feed

The two verticals are mounted in line with the direction most wanted. In other words, to work European stations from the East coast, the elements should be mounted in a northeast/southwest line.

Each element is fed with 52-ohm coax. Both pieces are cut to exactly the same length. However, a third piece, an electrical quarter-wave-length long, is inserted in series with one vertical, thus making it act as a reflector. This produces a significant gain in the forward direction and a very considerable front-to-back ratio. By using a relay or coax switch, this extra piece of coax can be switched from one vertical to the other, therefore instantly reversing the directivity of the array. Bill has found this feature to be very valuable and quite disconcerting to those on the other end of the contact.

## tuning

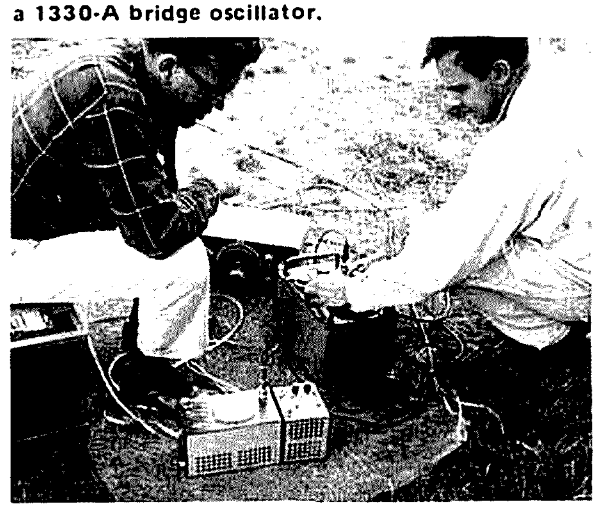
Tune-up of the array was checked with some General Radio test equipment, but for those with more meager means (almost all of us), each element can be tuned separately by using an swr bridge and adjusting the tap on the loading coil for minimum swr. This should be very close to 1:1 at the chosen operating frequency. The loading coils are mounted in small water-tight metal or plastic containers at the base of the elements.

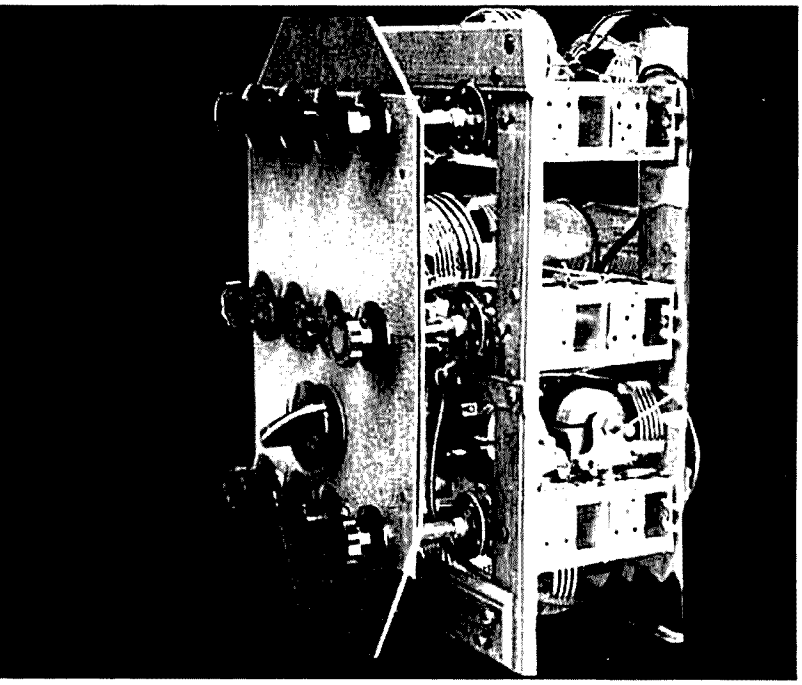
## results

Results have been excellent as might be expected. Front-to-back ratio is sufficient to knock down most of the W and VE interference. This, with the forward gain, gives a very significant boost to DX signals. While gain measurements are always tricky, it appears that gain in the order of 3 to 5 dB is reasonable, while actual on-the-air reports (compared with a well-tuned dipole) have been so good as to be considered a bit far-fetched. However, signals received with excellent strength on the other side of the world from VE1 can attest to the performance of the array.

ham radio

**Final check by VE1RK (left) and VE1AX. Test equipment is GR 1606-A rf bridge with 1212-A null detector and a 1330-A bridge oscillator.**





# antenna tuner for optimum power transfer

How to increase radiated and received power using the correct impedance transformer

George A. H. Bonadio, W2WLR, 373 East Avenue, Watertown, New York

In my article on the construction of diversity antennas,[1] I stressed the importance of using the proper antenna tuner. My antennas, because of their inherent low reactance, will work with almost any tuner. But to obtain maximum power transfer, and hence maximum efficiency, you should use what I'll call a "compatible" tuner.

A compatible tuner will transfer power from source to load regardless of load reactance. Conversely, an "incompatible" tuner, although it will transfer power, is inefficient because it doesn't account for reactance.

Both tuner types are discussed in this article. Data is presented to show you how to build a compatible antenna tuner—the most efficient deivce for transforming power from transmitter to load.

## an example

Suppose you wish to transfer power from a 52-ohm coaxial cable to a balanced load (fig. 1). The load might have inductive and capacitive reactances that exceed the resistive component (fig. 1B).

A whip antenna, for example, could have a resistive component of, say, 0.5 ohm (**fig. 2**). Its reactance could be 360 ohms

The important thing to remember is that if the ratio of reactance to resistance is high in the tuner, the compen-

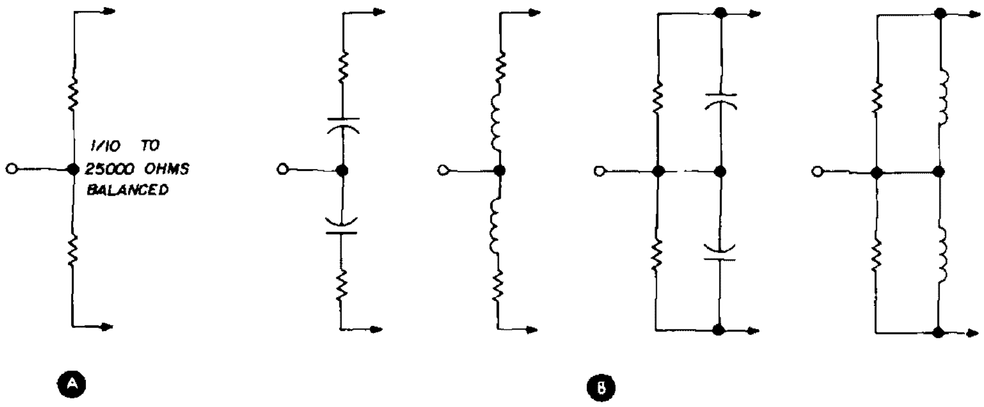



**fig. 1. The tuner problem. Load at A is resistive; a tuner working into it will transfer power with minimum loss. A tuner working into loads *in* B must compensate for reactances.**

capacitive. To obtain maximum power transfer, you'd need a reactance of opposite sign and equal magnitude. In this example a huge coil would be required, whose skin resistance might be 9.5 ohms. This would result in a 10-ohm series circuit consisting of 0.5 ohm radiation resistance (across which useful power is *dissipated*) *plus* 9.5 ohms (across which power is wasted). The circuit's efficiency would be only 5 percent.

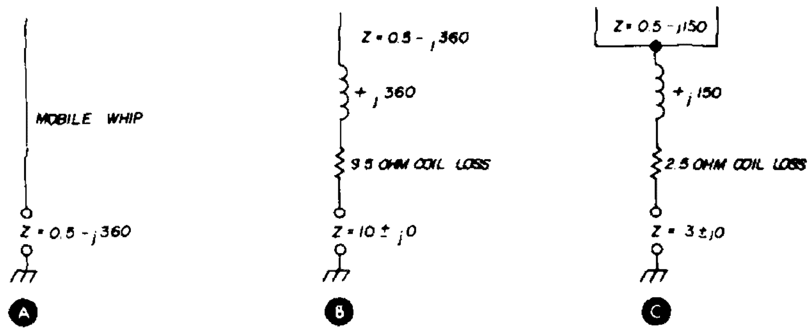
If a second whip antenna is added, as in **fig. 2C**, the capacitive reactance is

sating element may have to be physically large with consequent losses.

## power factor

A definition of power factor is the ratio of active (or true) power to reactive (apparent) power. It's generally expressed in percent. Your power company spends a lot of money in its generating equipment to overcome power-factor losses. A corollary in amateur work would be an antenna tuner that delivers only a fraction of the available

**fig. 2. Reactance compensation. Antennas must be resonant to accept power. This is shown in B, with 95% power loss in the coil. Antennas in C can provide 3 times as much radiated power.**




reduced to about one-half, and only 2.5 ohms of skin loss occurs in the compensating inductance. Three times as much power is then available for radiation. This idea might be extended to four balanced whips with interesting results.

power to the load. For example, I once had a tuner that delivered only ten percent of its input power to the load on some bands. This is a 90-percent loss, or 10 dB into and out of the tuner.

If the voltage peak doesn't agree with

the current peak in your tuning apparatus, a correction is needed. This is usually accomplished by tuning. If the tuner doesn't compensate properly for reactance, you've got problems.

## incompatible tuners

Perhaps you're using one of the tuners shown in the examples of **fig. 3.** It may work beautifully into a dummy load, but its efficiency will suffer from power-factor loss when working into a reactive load.

Consider the load resistance shown in **fig. 1A.** This is a purely resistive load; voltage and current are in phase, and most of the power is dissipated as useful energy. The loads shown in **fig. 1B** contain reactance. Unlike the power company, you don't have a reserve source of power to pump into the load—your generator, which is your final amplifier, is probably working at full output.

To compensate for the reactance, your final will have to supply, say, 100 volts at zero current; the resistive portion of the load will accept 100 volts at 1 ampere. The 100 volts at zero current is so-called "wattless power." Why accept a 3-dB reduction in your transmitted signal or the same loss of received signal power? The compatible tuner is the answer.

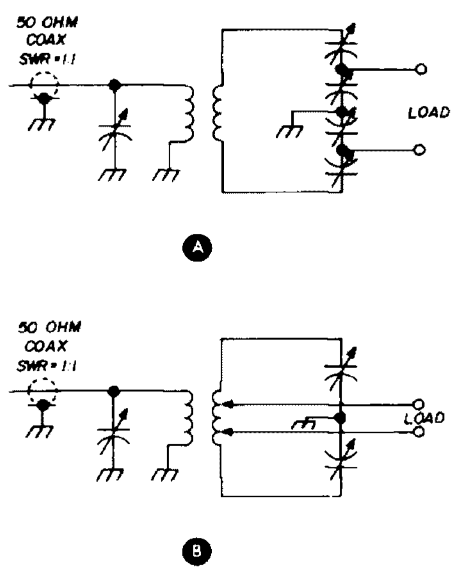



**fig. 3. "Incompatible" tuners. Regardless of unity swr, load reactances can cause power-factor loss that wastes energy.**

## compatible tuners

Don't be misled by a standing-wave ratio of 1. This doesn't prove you don't have a power-factor problem. An incompatible tuner can produce an swr of unity just about as well as a compatible tuner, regardless of load reactance. Neither tuner will *tune* differently, but a compatible tuner will deliver *full* power to its load **(fig. 4).**

## tapped constants

You can always tap down on tuner coils and capacitors. If the tuner feeds a resistive load, it will transfer power with

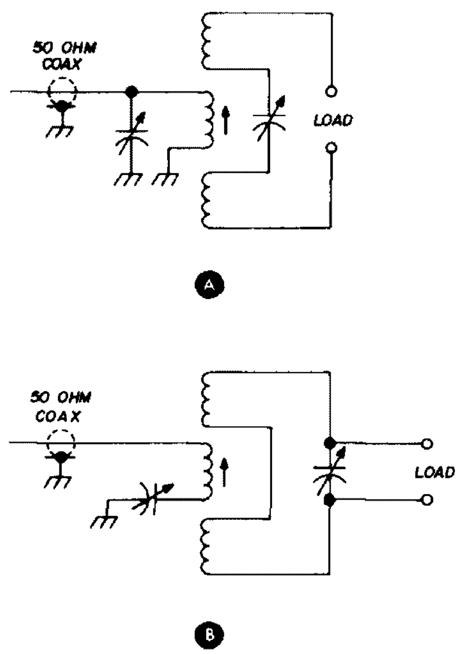



**fig. 4. "Compatible" tuners. Either will handle the load reactance. Note that feeders in B are connected directly to the capacitor.**

minimum loss. If, however, you tap across the tuner with a capacitor, power will circulate in the capacitor and will be returned to the load -sheer waste. The same reasoning applies to a tapped inductance. In general, there's an even chance that only 3 dB, or half your available power, will be delivered to the load if your tuner has the configuration of those in **fig. 3A** or **3B.**

In the compatible tuner, you always load across the lumped inductive and

capacitive reactances. No tapped elements are used. A tuner using differential capacitors, as in fig. 3A is electrically equivalent to tapping down on the inductance (fig. 3B).

## series solution

If line reactance and resistance are in series with a compatible tuner, all reactances cancel, and the system will be resonant. The series solution works fine if the tuner sees a balanced load of under 500 ohms. If the load is much higher, your coils will be so large you may want to change the design.

## parallel solution

Perhaps your load is higher than 500 ohms. In this case, you can connect a capacitor in parallel with the load impedance (fig. 4B). While this circuit has low peak voltage, it has high circulating current. A smaller coil will waste power because of the current. The changeover at 500 ohms is a matter of economy, physical space, and wavelength.

## practical circuits

To build a practical compatible tuner, measure line resistance and reactance, then plot the data. You can determine sizes of the constants from a table hookup. Then all you have to do is wind the coil, cut it and solder. It works right the first time.

## proof test

Disconnect your antenna feeders from the tuner. Connect a lamp in place of your antenna feeders (fig. 5A). Tune your transmitter into the load, then change bands. If you're using a compatible tuner, lamp brilliance won't change much from band-to-band.

Now connect the same lamp load as shown in fig. 5B. Excite the load, starting at the lowest-wavelength band, then progress through the shortest-wavelength band, If you're using an incompatible tuner, you'll find that the bulb won't light well on most bands.

In my next article, I'll show how to build a simple homemade tuner that can be used as an outboard device to compensate for reactance in an incompatible tuner. I'll also discuss feedpoint measurement using the substitution bridge methods.

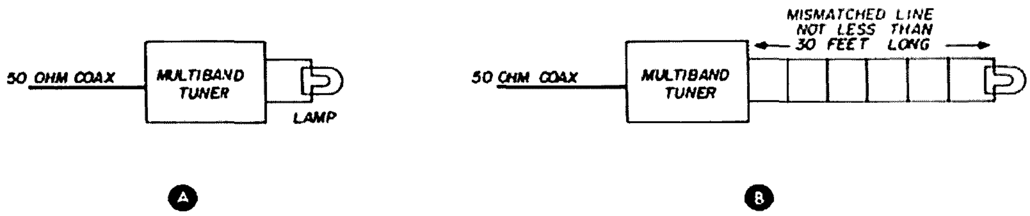
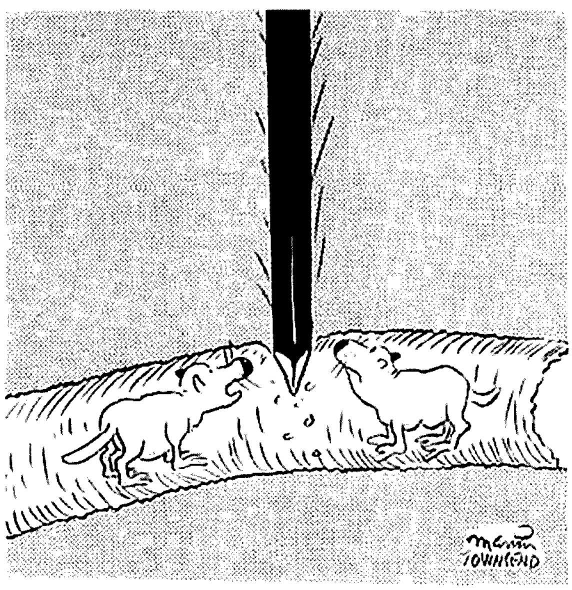



**fig. 5. Test circuits. "Compatible" tuner, A, will show same lamp brilliance on all bands. An "incompatible" tuner, with feed line connected as in B, will not deliver full power to the lamp on some bands.**


reference

1. George Bonadio, W2WLR, "Construction of High-Frequency Diversity Antennas," *ham radio,* October, 1969, p. 28.


**ham radio**



**"Watch your head! Our ham friend is putting in a new ground rod."**

# the isotropic source and practical antennas

Antenna performance is explained in terms of the imaginary free-space isotropic radiator

Ray Griese, K6FD

The isotropic antenna is an imaginary concept as basic to antenna theory as Ohm's law is to circuit theory. Over the years, Ohm's law has been expanded into many forms of circuit theory that are hardly recognizable to those with limited experience. But the most complex forms can be traced back, step by step, to E=IR. The relationship between the isotropic antenna and advanced antenna theory is similar. Unfortunately, very little has been written about the isotropic antenna on the basic level for use by radio amateurs. Perhaps this article will help those who desire a keener insight into operation and performance of antennas.

## antennas in free space

The isotropic antenna is considered as a point source of electromagnectic radiation. It is the simplest form an antenna can take. In fact, it is so simple no one has managed to build one for electromagnetic waves, although it has been approximated for sound waves. The difficulty is that a point-source antenna must have zero dimensions. Such an antenna, of course, is impossible to construct by any known method. But even so, this nice bit of fiction can be easily approximated for many practical purposes. For example, a 7-MHz antenna located 100 km in space will *look* like a point source even though it isn't. The errors introduced by the size of the antenna are too small to matter at that

distance for almost all areas of interest. In most calculations, a distance of 100 km isn't necessary to make an antenna look like a point source of radiation. This distance can be just a few wavelengths in some instances without introducing intolerable errors. The free-space radiation pattern of the isotropic antenna is shown in fig. 1. The antenna radiates in all directions: east, west, north, south, up and down. There is no inherent directivity, and this is exactly what the name

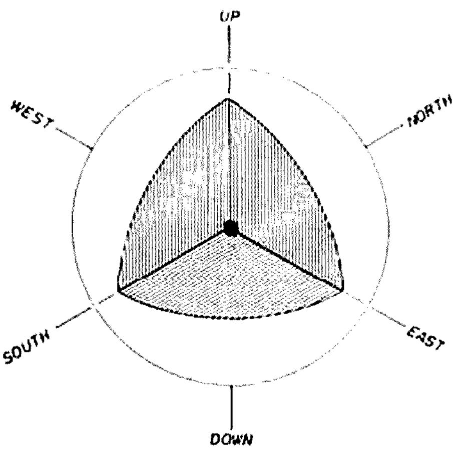



**fig. 1. Radiation pattern of the isotropic antenna in free space.**

implies: *iso* meaning everywhere the same; *tropic* meaning no matter how it is turned.

## power and voltage measurements

Antenna engineering involves measurements in addition to those of frequency, phase, resistance and resonance. Two of these measurements are power density in watts per square meter ($W/m^2$) and electrical field strength in volts per meter (V/m). Both can be derived by using the isotropic antenna for demonstration.

Imagine that the isotropic antenna is located in the center of a spherical balloon whose surface is everywhere 1 meter from the antenna (fig. 2) and that the antenna is radiating 1 kW. All of the radiated power must pass through the surface of the balloon, and it is easy to calculate how much power passes through each square meter of its surface The surface of a sphere is $4\pi R^2$. With a radius of 1 m, it is 12.57 $m^2$. Therefore, the power passing through 1 $m^2$ of surface is 1000/12.57, or 79.5 W. If the balloon is inflated until the radius is 10 m, the surface area will be increased 100 times, the area will be 1257 $m^2$. and the power passing through 1 $m^2$ will be 0.795 W. It can be seen that free-space loss in power, in its progress from source to oblivion, is due to the spreading of the wave. The mathematical relationship is $P = 1/D^2$, where D is the distance from source to point of measurement. Thus, the power passing through a unit area is inversely proportional to the square of the distance from the source.

In actual antenna-pattern and field-strength measurements, the instruments read out in voltage rather than power. The conversion from $W/m^2$ to V/m is related by

$$e = (377P)^{1/2}$$

where

e is V/m

P is $W/m^2$

377 is a constant*

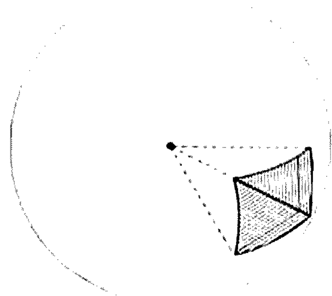


**fig. 2. Radiated power passing through 1 square meter of spherical surface.**

Returning to the spherical balloon example, if $W/m^2$ is converted to V/m, the field strength at D = 1 km is 0.173 V/m. This is usually expressed as 173 mV/m in antenna literature. It is the free-space field of 1 kW at 1 km from an isotropic source. Sometimes referred to as "unattenuated free-space field strength," it means that nothing is in the path between antenna and measuring equipment that

*The number 377 is the product of $120\pi$ which is the resistance of free space. Editor.

will introduce measurement errors.

## directional characteristics

Engineers use antenna directional characteristics to maximize radiation at the angles of elevation and azimuth required for the station's coverage. The principle uses vector addition of the radiated fields. This principle is well covered by antenna texts, so it will not be discussed further. But the use of the isotropic antenna in lieu of an actual antenna simplifies the problem to some degree, and it is helpful to look at it in this manner.

Suppose the isotropic antenna is ½ wavelength above a perfectly conducting and reflecting plane. Measurements of the radiated field will show that the spherical pattern has been distorted by the reflecting plane; the pattern is depicted in fig. **3**. Radiation upward is zero, because the reflected wave is reversed in phase and cancels the downward radiation. Radiation horizontally along the plane is also zero, because the reflected wave is reversed in phase, and the sum of the direct and reflected wave is zero (fig. **3**).

The path length of a wave reflected 30 degrees above the ground plane is ½ wavelength longer than that radiated directly from the isotropic antenna. The additional ½ wavelength adds 180 degrees of phase change. This, plus the 180-degree phase reversal due to reflection, brings the two waves back in phase, and they add. The sum is twice the value of either direct or reflected wave, resulting in 6-dB gain. If the wave from the isotropic antenna is horizontally polarized, then its pattern cross section is similar to that of a dipole located ½ wavelength above a perfectly conducting plane.

## the dipole

The directive property of the dipole is caused by wave cancellation and addition similar to that of the isotropic antenna above a perfectly conducting plane. Suppose the dipole consists of two isotropic antennas spaced ½ wavelength apart. At a distant point on the axis line, the radiation from one would arrive 180 degrees out of phase with the other, and would cancel the fields. At right angles to the axis line between the two isotropic antennas, the fields would arrive at the same time, in phase, and would add, thus making a lobe of maximum radiation. Minimum radiation, of course, would be along the axis of the line between the two isotropic antennas.

The field of a dipole is almost exactly like this. The difference is caused by power distribution along the antenna. If the dipole were cut into an infinite number of isotropic antennas whose power and phase were identical to that of the dipole, and the fields of all of these isotropic antennas were added, the result would be the field pattern of the dipole. Thus, the inherent directivity of the dipole exhibits a gain 2.1 dB above that of the isotropic antenna. The gain is not 6 dB, because it is generated in a different manner than that of the isotropic above the reflecting plane; it is the sum of many fields of many amplitudes and phases and not the sum of two fields of equal amplitude and identical phase. However, when the dipole is placed horizontally ½ wavelength above a perfectly conducting plane, the maximum lobe will be the sum of the direct and reflected fields, and the gain will be 6 dB. This gain, plus the dipoles's 2.1-dB inherent directivity will result in 8.1-dB gain above the isotropic antenna. This pattern is the familiar double-lobe pattern as shown in fig. **4**.

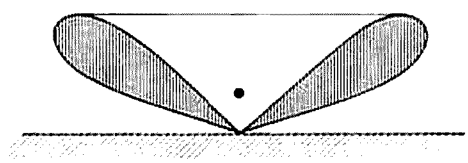


**fig. 3. Cross section of the radiation pattern from an isotropic antenna ½ wavelength above a perfectly reflecting plane.**

## the real world

A perfectly conducting and reflecting ground plane is hardly ever available. The average soil has much less than ideal

reflecting capabilities. Losses from this source distort the neat, geometric patterns shown in the textbooks.

Another interesting conclusion results from the reflection phenomenon. Suppose the antenna is erected over the world's worst location—the soil absorbs all the radiation, and reflection is zero. The field strength of the maximum lobe will be one-half the ideal value—a 6 dB loss. The minima will disappear, and the gain at those angles can be 10 to 20 dB, which may or may not be useful in transmission and reception.

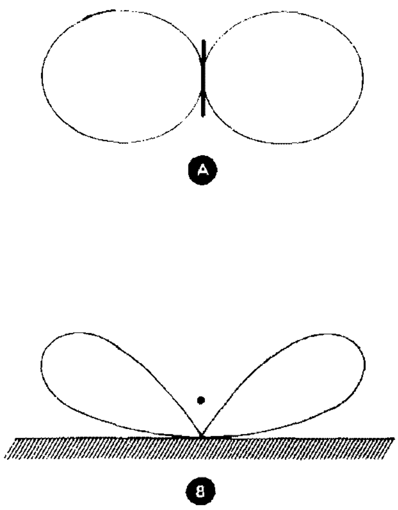


**fig. 4. Radiation pattern of a dipole located ½ wave above a perfectly reflecting plane; horizontal pattern is shown in A, vertical pattern in B.**

## grounded antennas

Grounded antennas also respond to the isotropic antenna concept. Suppose the isotropic antenna is located directly on the perfectly reflecting and conducting plane. The antenna is so close to the reflecting plane that no phase reversal of any kind exists. The pattern will be a hemisphere, because no radiation will exist downward through the reflecting plane. Assuming 1 kW is still being used, in free space 500 W will be radiated above the equatorial line in **fig. 1**, and 500 W below. When on the ground, 1 kW will be radiated upward, and the power gain will be 3 dB.

Vertical grounded antennas exhibit a modified 3-dB gain factor because of their finite length. Radiation from the bottom tends to generate the isotropic pattern, while that from the top tends to create maxima and minima, because the top is some distance above the reflecting plane. Grounded antennas shorter than 5/8 wavelength exhibit more gain toward the horizon than the sum of the 3-dB and free-space directivity gain. The extra gain comes from increased directivity at low vertical angles, because the wave is vertically polarized, and the direct and reflected waves add. There is no phase change with reflected vertical polarization.

Again considering the vertical antenna situated over the world's worst ground, maximum loss will be 3 dB! It is important to have the vertical antenna as far as possible from nearby structures.

It is also important to have a low-resistance ground connection. Vertical antennas usually have a feed-point resistance from a few ohms to a few hundred ohms. A ground-rod may have a ground contact resistance of 200 ohms; a 36-ohm antenna *connected* to such a ground would receive only about 1/6 of the delivered power—5/6 is used to heat the ground rod.

## radiation pattern and ground conductivity

Ground reflection varies with different soils. The amplitude and phase of the reflected wave depend on the angle of incidence, soil conductivity, and soil dielectric constant. The radiation pattern of grounded verticals in the high-frequency bands may be radically different from that anticipated. These verticals can be improved by using two or more ¼-wavelength ground-plane radials. On vhf bands and higher, vertical antennas are many wavelengths above ground, so the effects of soil conductivity and dielectric constant are minimized.* The maximum gain from a vertical antenna several wavelengths above a perfectly conducting

*However, a simulated ground plane, consisting of ¼-wavelength radials, is essential for vertical ¼ wave vhf antennas.**Editor.**

table 1. Directivity power gains of common antennas radiating 1 kW.

| antenna | power gain (numerical) | power gain (dB) | $\mu W/m^2$ at 1 km | $\mu W/m^2$ at 1 mile | mV/m at 1 mile | mV/m at 1 km |
|---|---|---|---|---|---|---|
| dipoles, free-space conditions | | | | | | |
| isotropic | 1.00 | 0 | 79.5 | 30.6 | 107.5 | 173 |
| 1/16 wave | 1.50 | 1.76 | 119 | 46 | 131.5 | 212 |
| 1/4 wave | 1.54 | 1.86 | 122 | 47 | 133 | 214 |
| 1/2 wave | 1.64 | 2.14 | 130 | 50.3 | 138 | 222 |
| 5/8 wave | 1.74 | 2.39 | 138 | 53.5 | 142 | 228 |
| monopoles, grounded, perfectly reflecting ground, vertical polarization | | | | | | |
| isotropic | 2.00 | 3.00 | 159 | 61.2 | 152 | 244 |
| 1/16 wave | 3.00 | 4.76 | 239 | 92 | 186 | 300 |
| 1/4 wave | 3.30 | 5.18 | 262 | 101 | 195 | 314 |
| 1/2 wave | 4.84 | 6.84 | 385 | 148 | 236 | 380 |
| 5/8 wave | 6.53 | 8.14 | 518 | 201 | 275 | 442 |
| dipoles, ½ wave above perfectly reflecting ground, horizontal polarization | | | | | | |
| isotropic | 4.00 | 6.00 | 318 | 122 | 215 | 346 |
| 1/16 wave | 6.00 | 7.76 | 477 | 184 | 263 | 424 |
| 1/4 wave | 6.12 | 7.86 | 487 | 188 | 267 | 429 |
| 1/2 wave | 6.53 | 8.14 | 520 | 201 | 276 | 443 |
| 5/8 wave | 6.92 | 8.39 | 552 | 213 | 283 | 456 |

ground is more than 6 dB. Gains for several antennas, modified by the ground plane, are given in table 1.

## physically small antennas

An interesing feature about antennas is that a small antenna can radiate a large amount of power, but physical limitations prevent this from being done efficiently. One limitation is antenna conductor resistance. As the antenna is shortened, radiation resistance decreases to a point where most of the power is dissipated in heat rather than radiation. Some large vlf antennas, for example, radiate only about 5 percent of the power delivered to them. On the hf bands efficient operation, say in excess of 50 percent, is possible by using a dipole that's small compared to a wavelength. The practical limit is in the vicinity of 1/16 wavelength. The radiation resistance will probably be between 0.5 and 1.0 ohm. With care, a 160-meter dipole can be constructed that is as small physically as a 20-meter dipole. A small, center-fed dipole is much more effective on the 2- and 3-MHz bands than a short, grounded vertical *without a low-resistance ground.*

## avoiding losses

Poor antenna performance is caused by pattern formation disturbed by improper construction, improper feeding, or distortion from nearby objects. Poor ground connections in vertical antennas don't apply to horizontal dipoles. The difficulty with the latter is due to losses at the element ends, which carry high voltage. The best way to avoid these losses is to use plastic guys or insulators with low leakage paths. The antenna should be carefully tuned and matched to its transmission line.

## conclusions

The isotropic antenna concept clearly shows why any practical antenna's field is highest near the antenna. For best results, the antenna should be as clear as possible from obstructions, and ohmic losses should be a small fraction of the antenna's radiation resistance. If all resistances that dissipate power uselessly are reduced to an absolute minimum, the amount of radiated energy can be increased by increasing antenna radiation resistance.

**ham radio**

# underground coaxial transmission lines

Underground coaxial transmission lines offer a number of advantages — here's the correct way to do it

Bob Ruyle, WØFCH, 420 Steinway Road, Lincoln, Nebraska 68505

Have you ever considered putting that coax lead-in underground? It has an aesthetic influence on the wife and the neighbors and if properly installed has some advantages over the old aerial installations. The number one item of consideration is advance planning. This is absolutely necessary to make sure you don't run into gas lines, telephone cables

Author Ruyle getting ready to install an underground cable run.

or power lines that may be buried in the area.

First draw a sketch of the exact location the cable will occupy and use it to check city drawings of the area before making the trench. Keep the sketch with your equipment manuals for future reference.

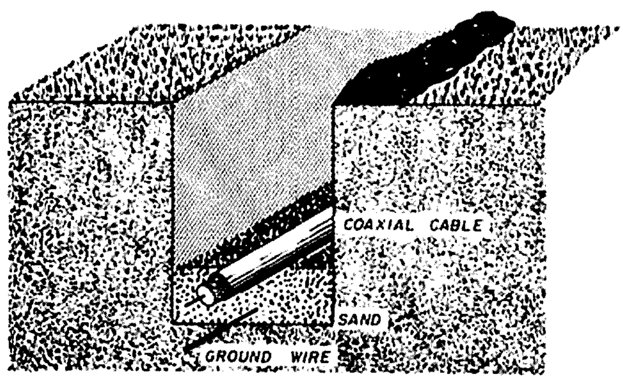



**fig. 1. Cable trench. Sand provides drainage and discourages rodents.**

## choosing cable

The next step is to choose the proper cable. Check manufacturers specifications for information on their cable; they will tell you which should be buried and which should not. The cable may or may not be armored. Armored cable should be used in areas where gophers and other rodents are a problem. However, when cable is installed as shown in **fig. 1** armored cable is not a necessity. Make sure you have enough cable to install it without a splice. (Splicing *can* be done, but great care must be taken to keep the moisture out.)

## the trench

Your next step is to dig the trench. Dig it 18 to 24 inches deep and 4 to 12 inches wide. If the terrain is rough increase the depth to 30 inches or more. Put 2 or 3 inches of sand in the bottom of the trench. (see fig. 2). The cable is placed on top of the sand. If you think you might need more cable in the future you can put a length of plastic conduit in the trench to allow for cable pulling at a later date.

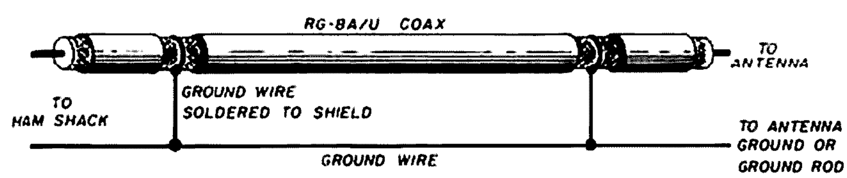
Buried cable is subject to lightning just as exposed cable is so a shield wire or a no. 4 AWG aluminum wire is recommended. Lay the wire beside or above the buried cable but not more than one foot away. The wire, just as the cable, ***should not be spliced.*** To be effective the shield wire must be connected to the outer shield of the coax cable at both ends as shown in fig. 2. (If the cable is spliced then the ground wire should be connected to the splice point as well.)

Sand is poured into the trench on top of the cable and the ground wire to a depth of 2 to 3 inches. The sand provides proper drainage to keep the cable dry and discourages underground rodents from chewing on the cable. The sand also protects the cable from rocks that may work their way up thru the soil.

Before refilling the trench be sure to check the coax for damage by attaching a dummy load to the end and running swr checks at spot points on each of the amateur bands. With a high-quality dummy load the swr should be 1:1.

ham radio

**fig. 2. External ground wire protects the coax against lightning damage. It shouldn't be more than a foot away from the coax.**



# integrated swr bridge and power meter

A novel application of a toroid as a sensor for a built-in swr and peak power monitor

Lloyd M. Jones, W6DOB, 17,779 Vierra Canyon, Salinas, California 93901

**The instrument described in this** article is designed to replace the various sizes and shapes of the black box known as the "swr bridge." The instrument, which I call an integrated swr bridge and power meter, performs all the functions of the usual swr bridge plus a few more—yet it occupies only about two cubic inches. Components will fit nicely in the space next to the rf output connector in most equipment. With the integrated swr bridge and power meter installed in your transmitter or transceiver, you can dispose of your regular swr bridge and use the parts for other projects.

The instrument is in the line at all times. It consumes only a few microwatts and doesn't disturb the circuit in which it's connected. In addition to measuring swr, the bridge measures peak power, monitors voice peaks, and indicates normal voice-power output. With a noninductive dummy load, you can check transmitter neutralization under full power.

The schematic is shown in **fig. 1**. The indicator can be your S-meter or plate-current meter. The meter sensitivity can be between 1 mA and 25 mA. A fixed resistor for R4, or a variable resistor controlled from the front panel, can be used to set peak meter reading to any desired value.

## circuit development

This circuit may be well-known to some readers, but I don't recall seeing it in this application. A schematic of a 50-MHz, 40-watt cw transistor transmitter in the RCA Transistor Manual[1] uses a vswr bridge to back- bias a driver stage for excitation control to the final amplifier. When the vswr is high, drive decreases to maintain the final amplifier transistor within ratings.

The RCA diagram, reproduced in **fig. 2**, suggests the possibility of adding another coil on the toroid. Thus the circuit could be used to indicate forward and reflected power with reasonable accuracy.

## construction

Toroid cores of one-half inch I.D. were used; a possible source is Amidon Associates.* The windings are made by twisting two lengths of No. 26 enamelled wire about 12 inches long. Wind the twisted pair to occupy ten turns around the core, then apply a coat of coil dope. Untwist about two inches from the ends, remove the insulation, identify each end with an ohmmeter, then connect the two windings in series (see **fig. 3**).

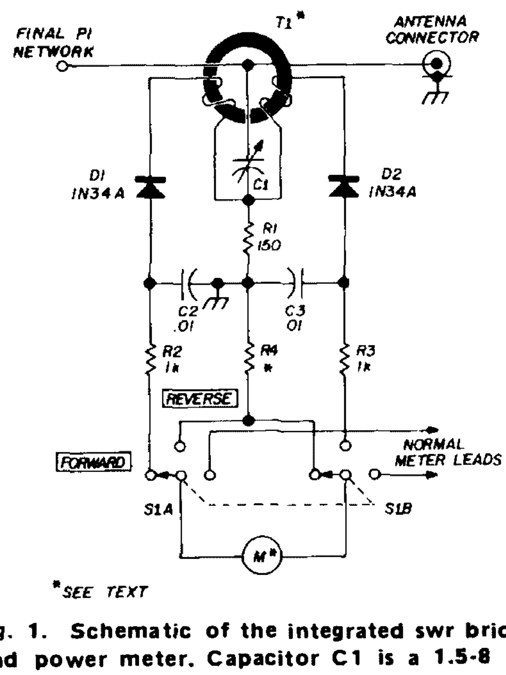
Unsolder the transmitter rf lead from the chassis connector and slip the toroid coil over the lead. Shim the toroid around the lead so that the coil can't possibly touch the rf lead. (I used a ½-inch length of polyethylene cut from RG-58/U coax and slipped it onto the rf lead; this provided a snug fit for the toroid.) Resolder the rf lead to the chassis connector. The lead from capacitor C1 (**fig. 1**) should be connected to the rf line within ½ inch of the toroid.

Cut a piece of phenolic board about 1 x 2 inches. A guide for laying out the parts is shown in **fig. 4**. If you'd like to go the modern route, etch the circuit as shown. Mount the parts as suggested in **fig. 4**. Bend the leads on the back side to join with other leads that are twisted together, cut off excess, and solder. Be sure to form solder connections close to the board, so that bushings about ¼ inch long will hold the board away from the mounting surface to avoid short circuits.

Attach three wires, preferably a 3-conductor shielded cable, to the board output. Route the cable to the meter and dress the cable to avoid the amplifier's rf field. Ground the cable shield at both ends.

The switch selects either forward or reflected power, or the *circuit normally* connected to the meter (marked A and B, **fig. 1**). If you don't wish to mount the switch on the panel of your equipment, the switch can be remotely located.

*Amidon Associates, 12033 Otsego Street, No. Hollywood, California 91607. Part no. T-80-2; price 60c each plus 25c handling; minimum order $1.00.




*SEE TEXT

**fig. 1. Schematic of the integrated swr bridge and power meter. Capacitor C1 is a 1.5-8 pF trimmer (Erie 539); S1 is a Mallory 22F134 wafer switch. Components occupy only two cubic inches.**

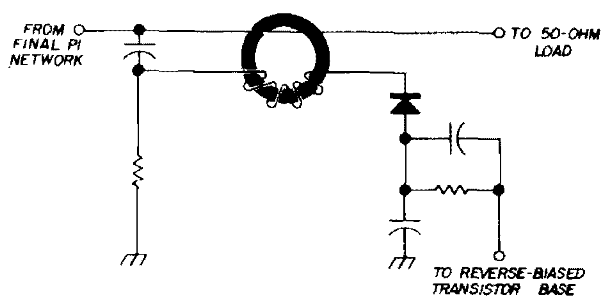



**fig. 2. Circuit in the RCA Transistor Manual from which the design was developed.**

## adjustment and calibration

When wiring is complete and checked, connect a dummy load to the transmitter. Adjust R4 and/or the carrier power to give a full-scale meter reading in the FORWARD position. Position the switch to REFLECTED. The reading should be noticeably lower.

When making adjustments, remember that high voltages are present in the vicinity of the circuit board. Keep clear of high voltage circuits and use an *insulated* tuning wand for all adjustments.

Adjust C1 for a minimum meter reading, which should be very close to zero. Check the adjustment of C1 on all bands. Capacitor C1 tunes out the reactance of

the toroid coil. If the dummy load is nonreactive on all bands, C1 shouldn't have to be changed with band switching.

Remove the dummy load and connect the antenna. Adjust the transmitter for full power, adjusting R4 as you tune to give a full-scale reading on the meter with the switch in the FORWARD position. Now switch to REFLECTED. The meter should read 1/10 full scale or less at the antenna resonant frequency if the antenna is matched to 50 ohms.

A rough indication of swr may be had by calibrating the meter (scale 0-100) as follows: 50 = 3:1, 25 = 1.5:1, 12.5 = 0.75:1, 6.25 = 0.37:1, and 3.1 = 0.19:1.

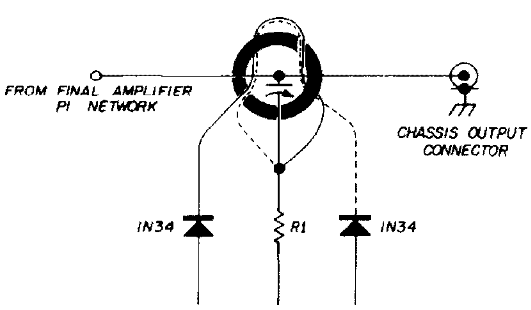



**fig. 3. Method of series connection for T1 windings. Only two turns are shown for clarity.**

If you adjusted C1 using a 50-ohm dummy load, then changed to a 75-ohm transmission line, C1 should be readjusted for minimum meter reading using a 75-ohm dummy load.

## peak-power measurement

As I mentioned at the beginning of this article, this instrument serves admirably as a peak-power measuring device. All that's required for this use is to choose a value for R4 for your particular transmitter input power (table 1). If your transmitter is rated at 180 watts input, for example, the value shown for an input of 200 watts would be adequate for R4. The output power will be some fraction of full-scale meter indication, of course.

If you decide to use a switch with another set of contacts, a fixed resistor could be used for R4 to indicate power output, and a variable resistor could be

**table 1. Values of resistor R4 for different transmitter powers and meter sensitivities.**

| transmitter input power (watts) | meter movement 1 mA | 500 μA |
|---|---|---|
| 100 | 7500 | 15k |
| 200 | 15k | 27k |
| 300 | 18k | 33k |
| 500 | 22k | 47k |
| 600 | 27k | 56k |
| 1000 | 47k | 91k |

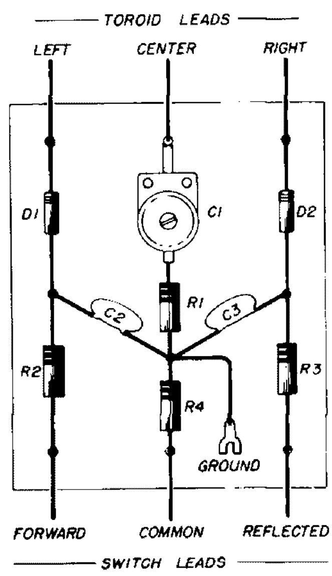
used to allow full-scale meter indication of swr with reduced power.

## conclusion

I've found this circuit indispensable for checking the resonance of my mobile whip antenna. With the switch in the REFLECTED position, I can adjust the whip antenna length for a minimum meter indication at the desired frequency.

Perhaps some of the more enterprising manufacturers of transmitters and transceivers will incorporate this instrument in future models. The small cost should make it worthwhile.

**fig. 4. Suggested parts layout on phenolic board.**




**references**

1. "RCA Transistor Manual," Radio Corporation of America, Electronic Components and Devices, Harrison, New Jersey, January, 1967 edition, pp. 504-505.

**ham radio**

# some reflections
## for
# reflected power

Transmission-line standing-wave ratio seems to be a topic for much debate — here are some interesting views on the subject

Walter H. Anderson, VE3AAZ

**A UA9 I worked recently said** he was using a Zepp antenna. It occurred to me that at least one generation has passed since the Zepp was, by far, the most popular ham antenna (see **fig. 1**). We didn't realize it then, but the Zepp's standing wave ratio probably ran as high as 30:1. However, history shows that the Zepp put out a good signal. Thus, it would seem that the Zepp didn't really have the side effects we hear attributed to high swr nowadays: high plate dissipation, radiation loss and all the rest.

I don't suggest we dismantle our beams and go back to Zepps. Rather, I propose to show that transmission-line theory, properly understood, is free of the contradictions that seem to arise when discussing swr, reflected power, line losses and other phenomena associated with antennas and feed systems.

### transmission-line length

The relationship between electrical and physical length is significant when discussing transmission-line characteristics. For example, the RG-8/U coax for

my 14-MHz antenna, is about 94 feet long. Its velocity factor is 2/3, which means the velocity of the signal traveling on the line is 2/3 of what it would be in free space. The line's electrical length is 94 x 3/2, or about 140 feet (two wavelengths). A signal requires $14.25 \times 10^{-8}$ second or 1/7 microsecond to travel the line's length (fig. 2).

Let's look at it another way. Suppose the letter H is sent at 12 wpm. In the time required to make just one dot, an electrical signal would make $3.5 \times 10^5$ round trips on the line. In this sense, the line is extremely short.

The transition of a conductor from a simple connecting wire to part of a transmission line is said to occur when circuit length exceeds 0.1 wavelength. Therefore most antenna transmission lines are electrically long and must be analyzed in terms of transmission-line theory.

In the first part of the discussion the line is assumed to have zero loss. Lossy lines are considered later from a practical viewpoint.

## line reflections

The characteristic impedance, $Z_0$, of a transmission line is the impedance which, at the line's termination, will absorb all the energy in the line. This is never realized in practice, because it's impossible to construct a line with constant impedance along its length.

What happens if the load impedance doesn't equal line impedance and some energy is reflected?

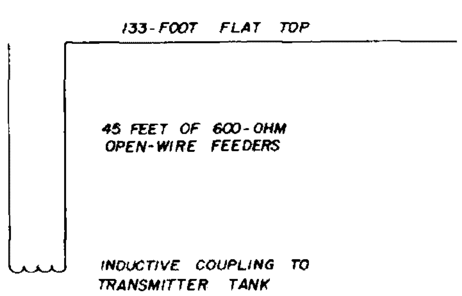



**fig. 1. The classic Zepp. A very popular antenna years ago despite a standing wave ratio that sometimes was quite high.**

It's unlikely that the source and line impedances will be exactly equal. Thus, any energy reflected from the load will travel down the line to the source to be reflected again toward the load. This repeats until the wave's amplitude becomes too small to be of interest.

The waves moving from source to load and from load to source are called the incident and reflected component, respectively. If the sum of any two waves is measured at any point on the line, the result would be a single standing wave. Thus, no matter how many individual waves are in the forward or reflected component, a standing-wave measurement will describe what's happening on the line at the point of measurement.

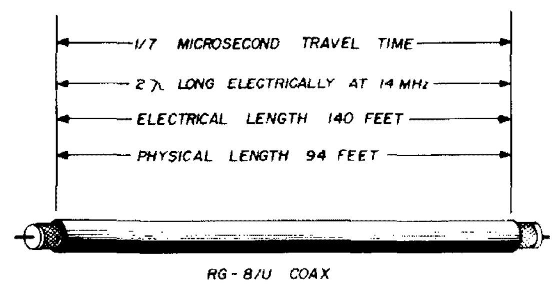



**fig. 2. Relationship between signal travel time and line length for a typical coax line.**

## standing wave ratio

Various instruments have been devised to sense the waves traveling from source to load and vice versa. The indicators (meters) on such instruments are calibrated for forward and reflected power. While these meters indicate in watts, we shouldn't regard the readings as we would usually use the word "watt." The only *real* power is the difference between forward and reflected indications on these instruments.

If a standing-wave pattern is made for voltage and some reflected power exists, the voltages will combine in phase at a voltage maximum and be out of phase at a voltage minimum. The ratio of these quantities is the voltage standing wave ratio. In a lossless line, the power passing

a given point will be the same as that passing any other point. Therefore, at points of voltage maximum, current will be minimum, and vice versa. It therefore

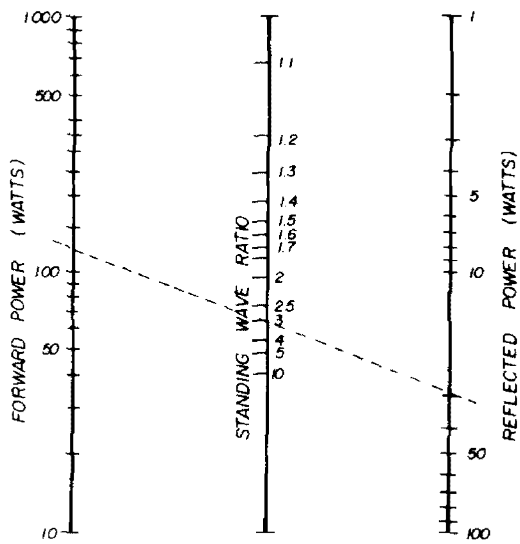



**fig. 3. Nomograph for finding swr for various values of forward and reflected power.**

follows that current and voltage swr can be used interchangeably.

Forward power, reflected power, and swr are related by

$$\text{swr} = \frac{1 + \sqrt{\frac{\text{Pref}}{\text{Pfwd}}}}{1 - \sqrt{\frac{\text{Pref}}{\text{Pfwd}}}} \qquad (1)$$

## common misconceptions

Up to this point the transmission line has been assumed lossless. This isn't true in practice although it's closely approximated in modern coax cable and open-wire lines operating below vhf.

Occasionally someone writes an article based on the graph shown in **fig. 4**, which was adapted from reference 1. The reasoning runs this way: if the terminating impedance is 100 ohms on a 50-ohm line, then the ratio of reflected-to-forward power is 1/9. So only 8/9 of the power is absorbed by the load, which means a reflection loss of 10 $\log_{10}$ 9/8 = 0.5 dB. The fallacious conclusion is that such a condition leads to a *reduction* in radiated power and possibly an increase in plate dissipation of the amplifier tubes. Viewed in proper perspective, this is insignificant in practical applications.

## lossy lines

Let's now use reflection loss in its proper sense. Consider a very lossy line with $Z_0$ = 50 ohms and a loss of 20 dB overall. If 60 volts are applied to the input, 1.2 amperes will flow. Thus 72 watts enter the line. At the load end the power will be down 20 dB, or down to 0.72 watt. If the load is, say, 100 ohms and reflected power is 1/9 forward power, the reflected power will be 0.08 watt, and only 0.64 watt will be absorbed by the load. The 0.08 watt will return to the source, suffering another 20-dB loss, arriving at the source at 0.0008 watt. This tiny amount of power will hardly affect conditions at the sending end.

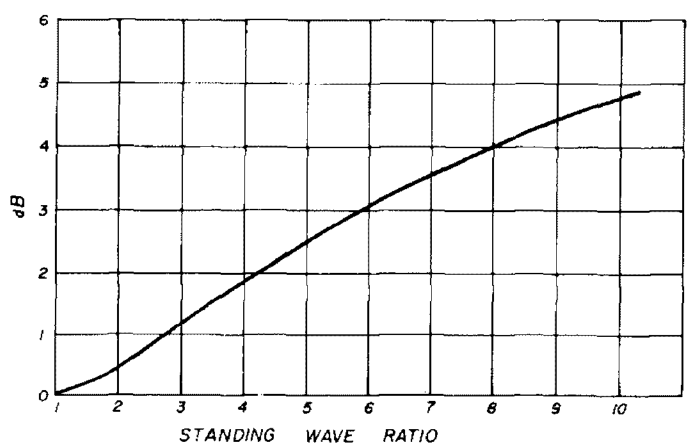



**fig. 4. Reflection loss on a transmission line with increasing swr.**

This is the justification for saying that the line looked like 50 ohms even though its termination was not 50 ohms. The significance is that the line loss is great; *not* that reflection loss exists.

Generalizing, we can say this about lossy lines:

**1.** The swr at sending and receiving

ends is different; the load end always exhibits the higher swr.

**2.** Power into the line equals the sum of line losses and power delivered to the load.

**3.** Reflection loss is only meaningful when line losses are so large that load changes don't modify the line's input impedance and thus cause a change in power into the line.

**4.** The lost power is not distributed evenly over the line, but is much greater nearest the source.

## approximate line loss

All lines have some loss. This is because standing waves of current stress the line at certain points. A rough guide of line loss as a function of swr is shown in fig. 5. An approximation of the loss can be made by shorting the line at the antenna, obtaining an swr reading and plugging it into the following equation:

$$\text{Line loss in dB} = \log_{10} \frac{\text{swr} + 1}{\text{swr} - 1} \qquad (2)$$

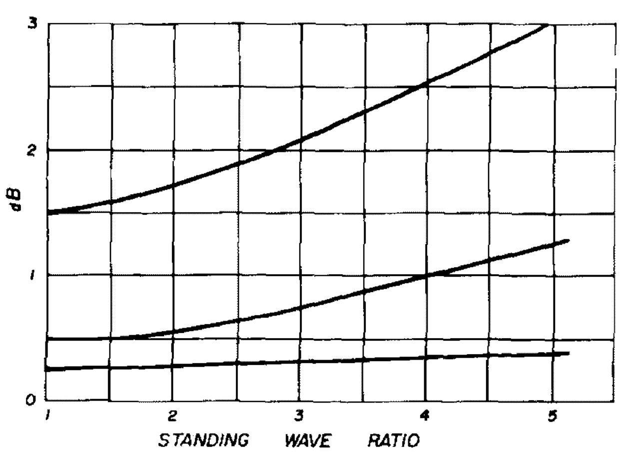



**fig. 5. Transmission line loss with increasing swr.**

## the Zepp again

Let's return for a moment to the good old days. Imagine a pair of 210's feeding a Zepp antenna that has a 30:1 standing wave ratio. Sixty watts would be a reasonable output for *100 watts input*, leaving 40 watts for plate dissipation and tank-circuit losses. The forward power would, however, be 480 watts, and the reflected power would be 420 watts. (Check the numbers in **Eq. 1** if you doubt it!)*

If there is no harm in high reflected power as far as *power delivered* to the antenna is concerned, then why all the fuss about high swr? We'll explore this and other fascinating topics next.

## transmitter specifications

Quoting from published data on two popular transmitters,

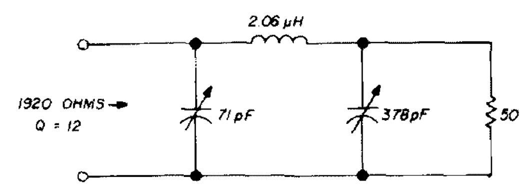



**fig. 6. Typical pi network values for 14 MHz.**

Collins 32S3: "Rf output impedance 50 ohms with not more than 2/1 swr."

B&W 6100: "Output impedence – will match 30-100 ohm resistive load."

These statements require some interpretation, since what is referred to in both cases is the *load* impedance, *not* the output impedance, which is what one would measure looking back into the transmitter output terminals. Also it's not a matter of matching (equalling or duplicating) anything; it's a matter of what impedance is appropriate as a load. As for the Collins, the statement *could* perhaps mean that as long as the resistive component was 50 ohms you could have up to 2/1 swr. What it undoubtedly *does* mean

*The forward and reflected power could also be 48 and 42 watts, or 4800 and 4200 watts respectively. Knowing the old 210 triode, 48 and 42 watts would probably be more realistic. **Editor.**

is that you can cope with any load impedance that has less than 2/1 swr on a *50-ohm basis*. As for the B&W, it appears that no reactance whatever is permissible, even though the swr can be about two or so. This suggests that the line must be trimmed to exact quarter-wave multiples.

If a low-pass filter is used, it would be difficult to meet the specifications without laboratory facilities. Perhaps these statements about loading are not intended to be taken this literally; however, the point remains—this is the way they are published.

## line sending-end impedances

When the reflected power is not zero (swr is not 1), and the line is of random length, the reflected voltages and currents may have any phase relationship when they combine with the forward voltages and currents at the sending end. Thus, the impedances at the transmitter end of the line may vary within certain limits, depending upon line electrical length. Table 1 lists some of these possibilities, all on a 50-ohm basis.

Table 1 shows that if you have a line with a known swr and put an rf bridge at the sending end, the impedance will fall in the corresponding column. The antenna's terminal impedance will also appear in the column. Note that the swr = 1 column has only one entry, while entries in the other columns are theoretically limitless. However, no impedance in any one column appears in any other.

So if you have a line with an swr greater than one, you can lengthen or shorten the line and change its impedance at the sending end. But you can never hit 50 ohms pure resistance, even though the *resistive component* may be 50 ohms plus some reactance.

Published transmitter specifications are undoubtedly intended to encourage the operator to keep the swr low so that the transmitter is presented with a load impedance near 50 ohms.

## output networks

Most modern transmitters use pi networks in the output. A common misconception is that these networks can transform any imaginable impedance to that required by the final amplifier plate circuit.

For example, consider a pi network designed to be used between parallel 6146's and a 50-ohm load. These tubes should work into a load of about 1920 ohms at a Q of about 12. The circuit is shown in **fig. 6**. The tank inductance is generally fixed, but the two capacitors are variable to maintain the proper load impedance despite external load changes.

What happens when the load is changed and the capacitors are adjusted to maintain correct tube plate impedance is shown in **fig. 7**. In this case, we face an impossible task when loads *are lower than* 17 ohms or higher than 1800 ohms. Also,

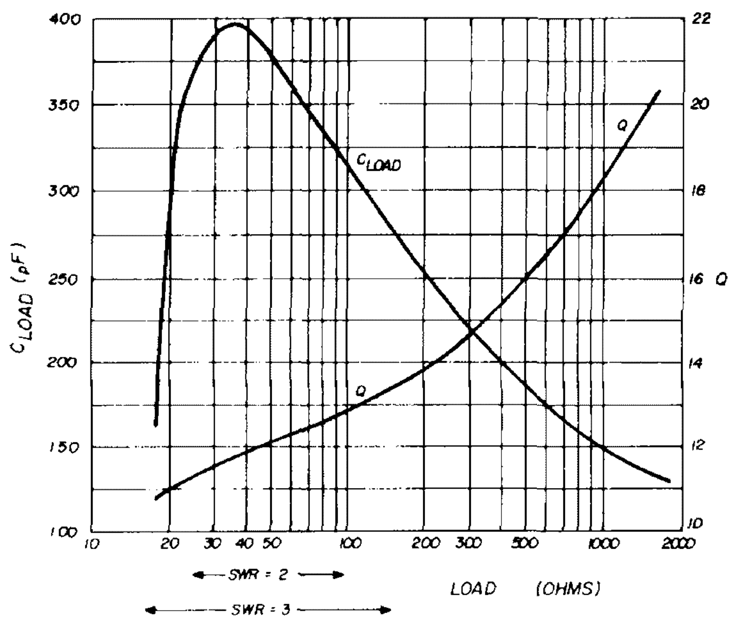



**fig. 7. Pi network loading capacitance and Q as functions of output loading resistance.**

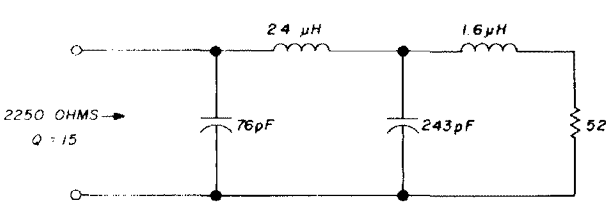



**fig. 8. Typical pi-L network circuit values for 14 MHz.**

even when we're well within the range of possible impedances, the loading capacitor must be varied widely, which is certain to influence harmonic response. Circuit Q also changes, and this will affect the output waveform, although probably not seriously. The only way to avoid this is to present the transmitter pi network with an impedance close to 50 ohms, i.e., a load with an swr close to unity.

When I was writing this article, Bill

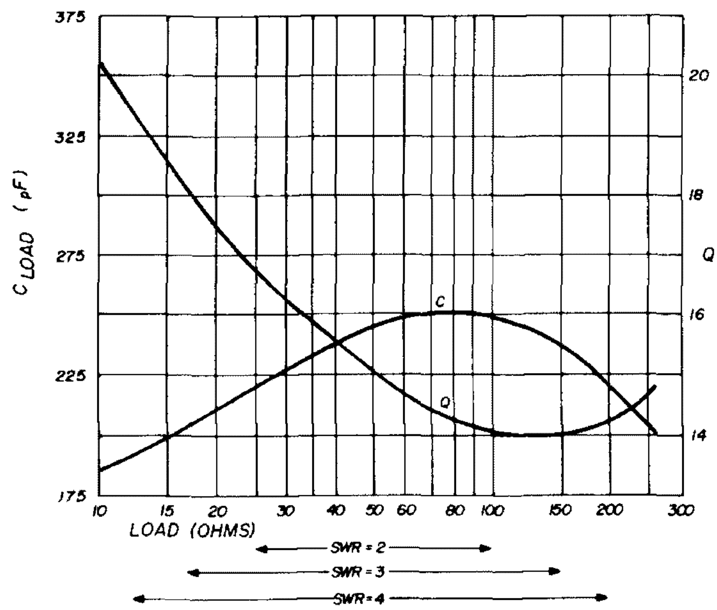



**fig. 9. Changes in pi-L network parameters for various loads.**

Orr's article on pi an pi-L networds came along.[2] Taking these parameters (input impedances = 2250 ohms; Q = 15; load impedance = 52 ohms) at 14 MHz, fig. 8, a similar investigation was made. The results, fig. 9, indicate that the pi-L will work somewhat more smoothly over a wider range. However, a limit still exists for the load impedances that can be handled; 10 ohms is now the lower limit.

To sum up,

1. Low reflected power means low swr.
2. An swr close to one means impedances close to 50 ohms.
3. Impedances close to 50 ohms mean that the transmitter can operate near its design center.

## antenna relays

Let's now go back to table 1 and compute some currents and voltages on the basis of 1 kW being sent down the line. This is depicted in **table 2**. At the higher swr's the "average" currents and voltages will be higher. This, of course, means greater line losses, as described earlier.

Most modern stations use some form of antenna changeover device for vox or fast cw break-in. Consider the antenna relay. Assuming worse-case conditions, the relay could be in the line at the point of maximum current. If the relay is rated at "1 kW for swr = 1," it's really rated at 4.48 amperes of contact current (**table 2**). If the number in tables **1** and **2** for resistance and current are substituted in the equation for power, you'll find that the power presented to the relay is 1 kW for an swr of either 1 or 2. However, the maximum current at an swr of 2 is 6.32 amperes. So if a relay with a maximum contact-current rating of 4.48 amperes is operated at the current maximum point in the line with a 2:1 swr, the relay's true *power* rating will be $(4.48)^2 \times 25$ ohms, or 500 watts. Clearly, such a "power rated" relay should be derated for maximum power-handling capability of

**table 1. Possible line impedances for standing wave ratios between 1 and 2.**

| swr | 1 | 1.5 | 2.0 |
|---|---|---|---|
| $\frac{Pref}{pfwd}$ | 0 | $\frac{1}{25}$ | $\frac{1}{9}$ |
| possible impedances | 50 | 60 + j20 | 75 + j35 |
| | | 60 - j20 | 75 – j35 |
| | | 33.333 | 25 |
| | | 37.5 + j12.5 | 40 + j30 |
| | | 375 - j12.5 | 40 – j30 |
| | | 75 | 100 |
| | | 50 + j20.4 | 50 + j35.3 |
| | | 50 – j20.4 | 50 – j35.3 |

**note: the highest purely resistive impedance is 50 x swr; the lowest purely resistive impedance is 50/swr.**

table 2. Voltage and current as a function of swr with forward power = 1000 watts.

| swr | 1 | 1.5 | 2 |
|---|---|---|---|
| $I_{max}$ (amps) | 4.48 | 5.48 | 6.32 |
| $I_{min}$ (amps) | 4.48 | 3.65 | 3.16 |
| $V_{max}$ (volts) | 224 | 274 | 316 |
| $V_{min}$ (volts) | 224 | 183 | 158 |

1000/swr watts. This is, in itself, a good argument for a low swr.

## antenna tuners

It's easy to dispense advice on obtaining a low swr, but it's much more difficult to specify remedies for curing high swr's. If you must live with kinky antenna impedances, then you might consider using an antenna tuner. If air dielectric capacitors and silver-plated inductors are used, power loss from the tuner will be negligible. An antenna tuner will allow the impedance presented to the transmitter to be close to 50 ohms, and transmitter specifications will be satisfied. Such a tuner also pays dividends when receiving.

In **fig. 10** the antenna is assumed to represent a load resistance of 25 ohms (swr of 2). The antenna is connected to either transmitter or antenna tuner through a lossless line two wavelengths long.

Suppose we accept the idea of reciprocity, so that the same antenna will have a source resistance of 25 ohms when receiving. Also assume that the receiver's input resistance is 50 ohms (actually, in many receivers input resistance is close to 300 ohms).

In **figs. 10A** and **10C**, where no tuner is involved, line swr will be 2 and 1 respectively. In **fig. 10B** the tuner will transform the load so that 50 ohms is presented to the transmitter, and the short length of line between tuner and transmitter will have an swr of 1.

If tuner settings are unchanged and the receiver is switched in **(fig. 10D)**, the swr on the main transmission line will increase to 2:1! What may be even more difficult to accept is that the power delivered to the receiver will *increase*—not much, true, but an increase nevertheless.

The reason that *more* power goes down the line in **fig. 10D** than in **fig. 10C** is because an impedance match exists at the antenna terminals in **fig. 10D**.

## conclusions

Transmission lines with low inherent losses should be used. Disregard reflected power as *power* and think of it as contributing to higher standing wave ratios.

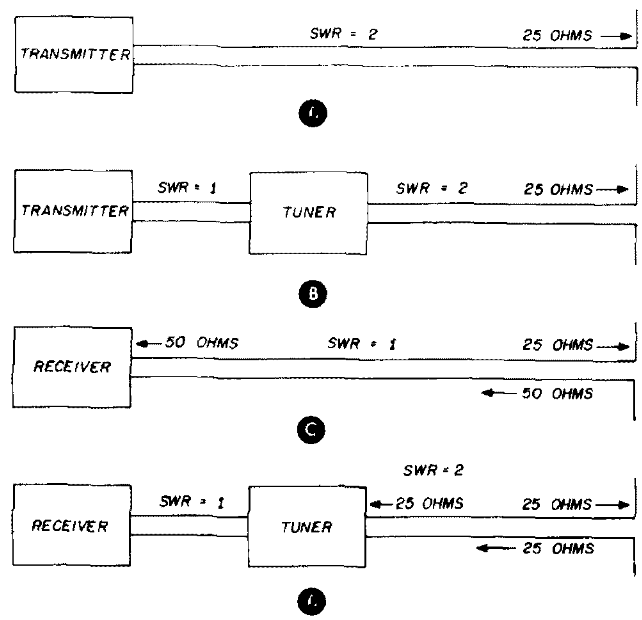



**fig. 10. Examples of what happens when using an antenna tuner for transmitting and receiving. Line impedance is 50 ohms in all cases; line is 2 wavelengths long.**

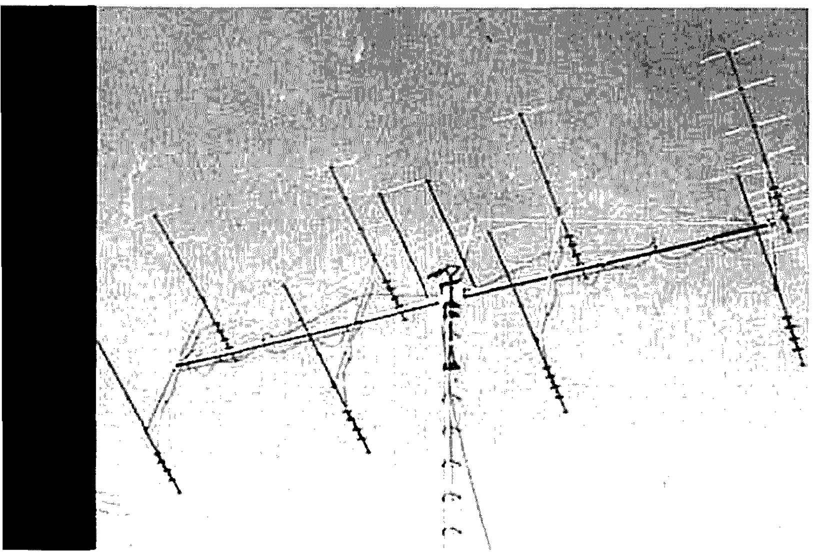
An antenna tuner is worthwhile for both transmitting and receiving, especially when transmitting swr is high.

Beware of claims by electronic equipment manufacturers. Power-rated devices for antenna changeover functions and published specifications for transceiver or transmitter output swr requirements can be misleading.

### references


1. "Networks, Lines, and Fields," Ryder Prentice-Hall, 2nd edition, p. 303.
2. William I. Orr, W6SAI, "Pi and Pi-L Networks for Linear Amplifier Service," *ham radio*, November, 1968, pp. 36-39.


ham radio



photos by W6BUR

# a practical 144-MHz moonbounce antenna

Here's a practical moonbouncing antenna that provides outstanding performance for tropo contacts as well

Ken Holladay, K6HCP, 7733 Rainbow Drive, San Jose, California

Earlier this year a radical new antenna design—the log-periodic yagi or LPY—was introduced to vhf amateurs[1,2]. This new antenna has some interesting characteristics that should interest any vhf enthusiast: broad bandwidth and relatively high gain. The 9-element two-meter Swan log-periodic Yagi* features 4-MHz bandwidth and 11.5 dB gain. And, unlike some commercial antennas, these characteristics hold from one antenna to another.

Both broad bandwidth and high gain are important for long-haul tropo work or moonbounce. Four of the Swan LPY antennas will give 17- to 18-dB gain, and eight of them are capable of successful moonbouncing. The array of eight Swan LPY antennas described in this article is being used for serious moonbounce work, and is a tremendous performer over long-haul tropo paths as well.

*Available from Swan Antenna Company, 646 North Union Street, Stockton, California 95205.

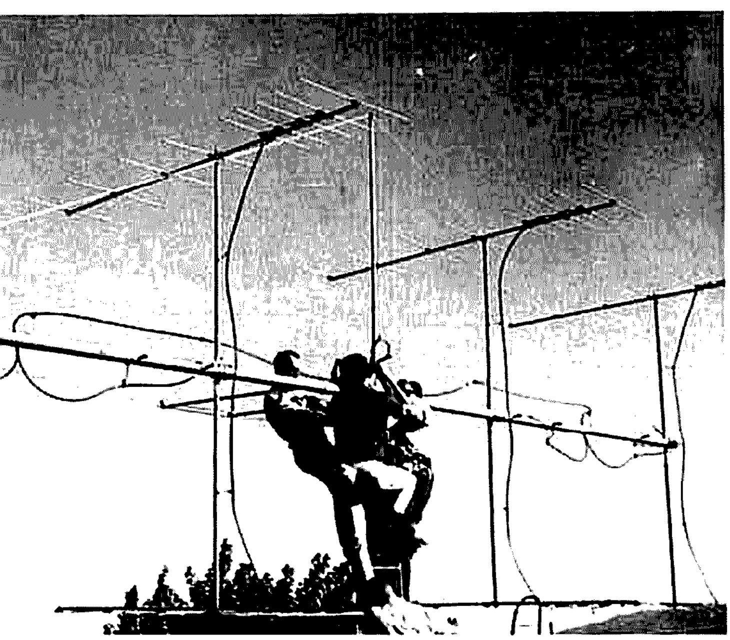
## construction

I am not going to describe the mechanical portions of the antenna in any great detail because locally available materials will pretty much dictate how you build your own array. Briefly, however, this antenna uses a 30-foot section of 3-inch aluminum irrigation pipe for the main boom. (4-inch aluminum would be better if it is available.) The LPY supports are made from 10-foot sections of 1-inch diameter tv masting.

The elevation system is made from two pieces of 3/8-inch thick aluminum plate, hinged at the bottom. The 3-inch boom is mounted on the side of the plate and the tower mast is mounted on the other. A worm gear and motor drive lowers and raises the plate, changing the vertical elevation angle of the array. I'm not going to provide exact details of the elevation drive because it was built up from surplus parts as we went along, more devised than designed.

Be sure to keep the boom from sagging. I used nylon rope for my installation, and it seems to be satisfactory. *Do not* use hemp; it shrinks when wet.

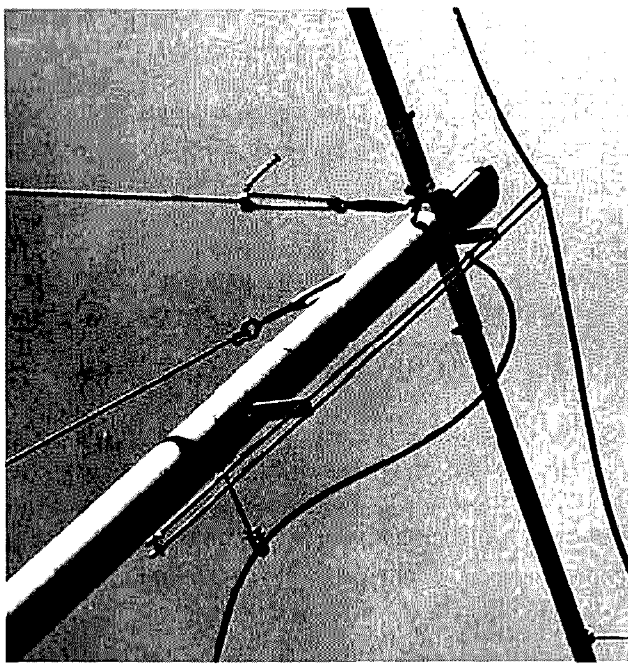
**There's not much room at the top the tower. Left to right: K6MYC, WA6MIA and K6HCP; WA6UAP below.**



## feeding it

The feedpoint impedance of the Swan LPY antennas is 110 ohms. This is not too convenient for 50-ohm coax, but the matching system shown in fig. 1 works well. High quality 300-ohm feedline is used throughout the matching system. (Belden 8275 is highly recommended.) All the lines are cut to multiples of one-half wave-length; the exact electrical length is determined by grid dipping a

**Corrective stub for feeding two of the LPY antennas.**



half-wavelength section and using it as a standard for the half-wave multiples. This is the method described by W1HDQ in the ARRL "Radio Amateurs VHF Manual" (page 182).

I must stress the importance of using half-wave multiples in the feed system. When this antenna was first put up half-wave multiple feedlines were not used and matching was very difficult; in addition, the lines were heavily influenced by surrounding objects. The half-wave multiple lines eliminated these undesirable effects.

Each pair of LPY antennas is connected through a 1-wavelength section of line to a corrective matching stub; these matching stubs are adjusted to match a 300-ohm line. Each *pair* of matching stubs is then connected through a half-wavelength line to a common standoff point. These two junctions are connected through 2-wavelength lines to another corrective stub, located at the center of the array. This latter stub is adjusted to match the characteristic impedance of the feedline you're using. My array is fed with 300-ohm Belden 8275.

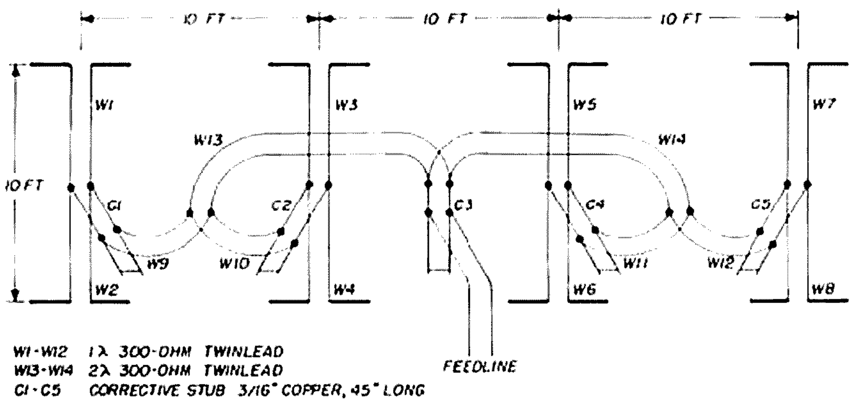



**fig. 1. Matching feed harness for the practical 144-MHz moonbounce antenna.**

A balun must be used for adjusting each matching stub. I used the one described in *QST*[3] (also in the ARRL vhf handbook, page 183). Don't use the flexible coax type because they are not very predictable. For adjusting the stubs on my array I used a 50- to 300-ohm balun, an inexpensive low-power swr meter and a Gonset Communicator at 144 MHz. The swr meter does not read true swr accurately at 144 MHz, but it does null when terminated with a 50-ohm load.

The matching stubs are 45-inches long and made from 3/16-inch hard-drawn copper that I found at one of the local scrap yards. If you don't have any material for these stubs, you can buy aluminum ones from Cush-Craft*. If you don't want to put up an array as large as eight LPY antennas, the matching scheme will also work for two or four LPYs.

## performance

On-the-air performance of the 8-LPY array has turned out just as expected. In the first two weeks of operation, K6MYC and I ran extensive tests and found that this array is 3 to 4 dB below his colinear.[4] I can hear my own echos off the moon, as well as those of K6MYC's. I have worked W1FZJ/KP4, almost worked SM7BAE, and my moonbounce signals have been heard in Australia.

**K6MYC inspects the elevation system.**

*Cush-Craft, 621 Hayward Street, Manchester, New Hampshire 03103.

For those of you who would like to try 144-MHz EME, I think this is the most practical antenna to come along. If you're a little more daring, an array of 16 LPY antennas should be really tremendous.

I would like to thank Ed, WA6MIA, Pat, WA6UAP, George, W6BUR, Glen, WB6VYM, and especially Mike, K6MYC—without his enthusiasm and mechanical talents this array would never have gotten off the ground.

K6MYC makes the final adjustments.

## bibliography


1. W. Orr, W6SAI, D. Crowell, K6RIL, "Log-Periodic Yagi Antenna," *ham radio*, July, 1969, p. 8.
2. E. Tilton, W1HDQ, "The Swan Multidrive 2-Meter Antenna," *QST*, October, 1969, p. 42.
3. K. Holladay, K6HCP, D. Rarwell, WA6GYD, "Beer-Can Baluns for 144, 220 and 432 MHz," *QST*, February, 1965.
4. "World Above 50 MHz," *QST*, January, February and March, 1968.


ham radio

# receiving antennas

A discussion of special purpose antennas for receiving – including some novel ideas for improved performance on the lower bands

Robert L. Nelson, K6ZGQ

It has become popular among amateurs, commercial and military stations to use the same antenna for receiving as that used for transmitting. In fact, amateurs have gotten so hooked on the idea that we have completely overlooked the tremendous advantages that specially-designed receiving antennas can offer. Take the forty-meter phone band as an example: the difficulty is not so much in getting your signal out as it is trying to dig the other station's signal out of the QRM. Wouldn't it be nice to have a receiving antenna that could null out the interference while simultaneously maintaining reception of the desired station? Sure, you *can* build a full-size Yagi and stick it up 70 feet in the air, but how many amateurs have that kind of money? Or room?

On forty meters you can probably work almost anybody you can hear with just a little power even if you use a simple vertical (or dipole) antenna. Why not use the vertical for transmitting and build a small, easily rotatable, *receiving* antenna that can make listening to forty a lot more enjoyable?

## the receiving system

The trouble with most antenna books is that they talk only about antennas. That seems reasonable enough until you

realize that the important characteristics of a receiving antenna can't be specified without considering both the antenna and the entire receiving system. You have to look at receiver noise as well as the electrical noise that the antenna sees.

A basic receiving system is shown in fig. 1. It consists of an antenna, a transmission line and a receiver. The antenna acts as a transducer between the impinging electromagnetic waves, (containing both signal and noise) and one end of the transmission line; the transmission line provides a path to the receiver. The receiver amplifies and detects both signal and noise, adds some noise of its own and presents the result to the listener.

The basic receiving system problem is to maximize the ratio of signal power to noise power at the receiver output. SNR, the output signal-to-noise power ratio is defined thus:

$$SNR = \frac{S_o}{N_o} \tag{1}$$

where $S_o$ is the available output signal power and $N_o$ is the available output noise power. If we had a perfect antenna, transmission line and receiver SNR would be the same as the ratio of the signal and noise powers as they propagate through space. This is the ultimate SNR, and you can't do any better.

There are three basic types of noise that make up the noise contribution to this ultimate SNR: atmospheric noise, cosmic noise and man-made noise. Atmospheric noise is generated by the electrical discharges that attend the hundreds of thunderstorms going on around the world. In the high-frequency portion of the spectrum (3-30 MHz) this noise energy is propagated by the ionosphere and is more evident below 30 MHz.

Cosmic noise is generated by the stars, and most reaches the earth from staggering distances; cosmic noise tends to dominate the noise picture between 20 and 200 MHz.

Man-made noise is just what the name implies. It represents the rawest form of pollution of the electromagnetic spectrum. Even in quiet locations this noise is often dominant over others in the region from 10 to 100 MHz, although these figures are highly variable; in less fortunate locations man-made noise may completely cover the high-frequency spectrum. Typical man-made noise generators are electric motors, neon lights and insufficiently shielded rf generation equipment.

You can't improve the ultimate SNR, but let's see what happens when we add a lossy transmission line or a noisy receiver to the system. The *free-space* signal and noise are attenuated equally as they pass through the lossy transmission line; thus the SNR at the receiver end of the line is identical to the ultimate SNR. However,

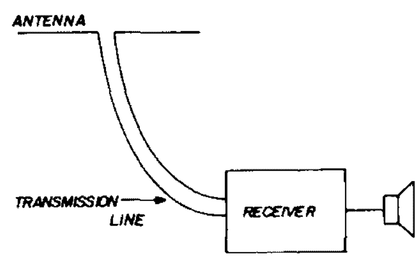



**fig. 1. Basic receiving system consists of an antenna, the transmission line and a receiver.**

at this point noise is contributed by the receiver; the result at the receiver output is a degraded SNR.

## antenna theory

At this point let's review some basic antenna theory and define directive gain, power gain, radiation efficiency and capture area of an antenna. Reference 1 gives an even more simplified treatment of basic antenna theory if you need it.

**Power gain.** Most simply this can be defined, in the receiving sense, as the ratio of signal power received from a distant station with a directive array, divided by the power received with a perfectly non-directional (i.e., isotropic) antenna. *It has the symbol* $G_p$.

**Directive gain,** $G_d$, is a measure of the directivity of an antenna. It can be

thought of most simply as a measure of the *shape* of the antenna radiation pattern. The radiation patterns of two different antennas are shown in **fig. 2.** Since the shape (but not size) of their radiation patterns is identical, they both have the same directive gain. But the power gain of antenna A is four times that of antenna B. We'll come back to this in a moment.

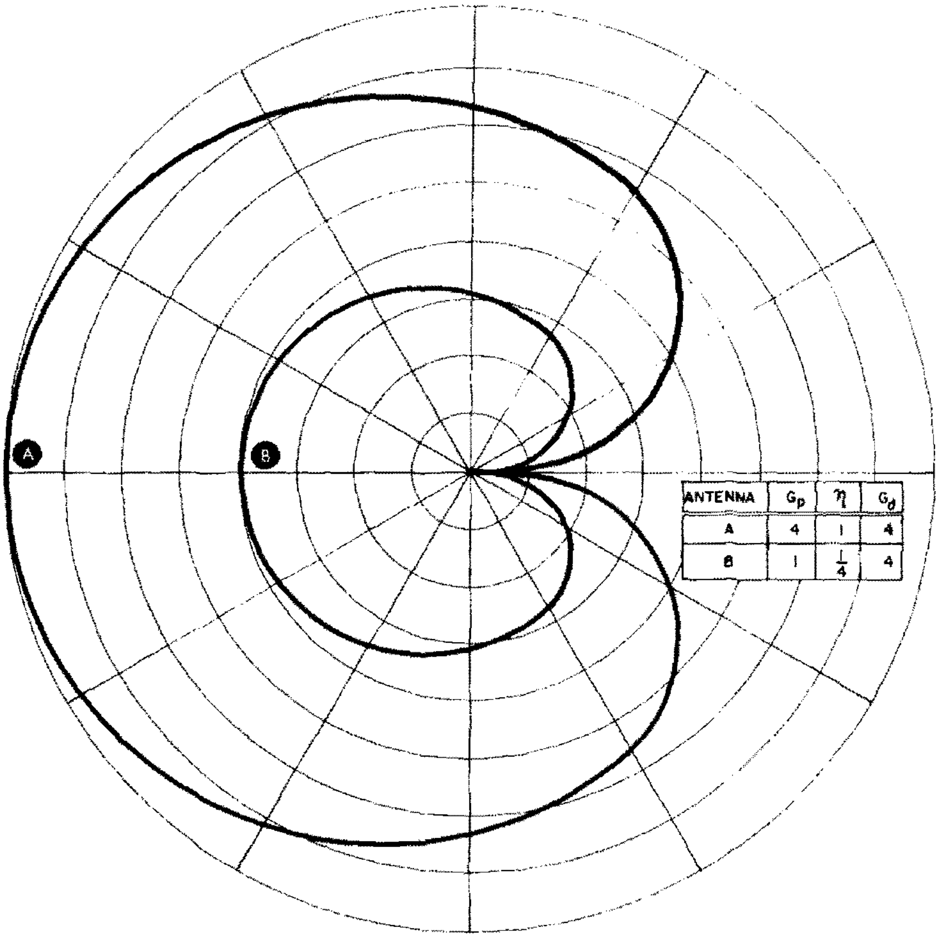
Every antenna has some losses. That is, if the antenna is used for transmitting, part of the transmitter output power will be burned up in heat losses due to resistance in the antenna. When the antenna is used for receiving, a portion of the received signal (and noise) power is burned up in the same lossy parts. These power losses are due to resistive elements present in the antenna (such as the resistance of wire, tubing and traps) and also due to ground losses and the like in the environment surrounding the antenna. The cumulative effect of these losses can be lumped into *loss resistance,* $R_l$. The useful signal power (transmitted or received) can be represented as resistance $R_r$; this is the radiation resistance. The resistive equivalent circuit of a resonant antenna consists of $R_l$ in series with $R_r$. The *radiation efficiency* of an antenna is defined as:

$$\eta = \frac{R_r}{R_r + R_l} \qquad (2)$$

where $\eta$ is the symbol for radiation efficiency. If the transmitter output power is multiplied by $\eta$, the result is the amount of useful radiated signal power. Power gain and directive gain are related by:

$$G_p = \eta G_d \qquad (3)$$

This accounts for the size difference in




| ANTENNA | $G_p$ | $\eta$ | $G_d$ |
|---|---|---|---|
| A | 4 | 1 | 4 |
| B | 1 | $\frac{1}{4}$ | 4 |

**fig. 2. Comparison of two antennas with identical directive gains.**

the radiation patterns of antennas A and B in **fig. 2**.

Now we come to "capture area." If an impinging electromagnetic wave has a power density $D_p$, then the power from the wave which is available at the antenna terminals is:

$$P_{av} = AD_p \qquad (4)$$

where $P_{av}$ is the available power and A is the antenna capture area. Capture area is the effective frontal area which the antenna presents to the passing electro-

put on a number on how much we can reduce antenna efficiency for a given degradation of SNR. For example, if the antenna efficiency is reduced until the system output SNR is 1 dB less than the ultimate SNR, the minimum required $\eta$ is:

$$\eta = \frac{T_o + LT_r}{0.26T_a + T_o} \qquad (6)$$

in decibels:

$$\eta db = 10 \log_{10}(T_o + LT_r) - 10 \log_{10}(0.26T_a + T_o) \qquad (7)$$

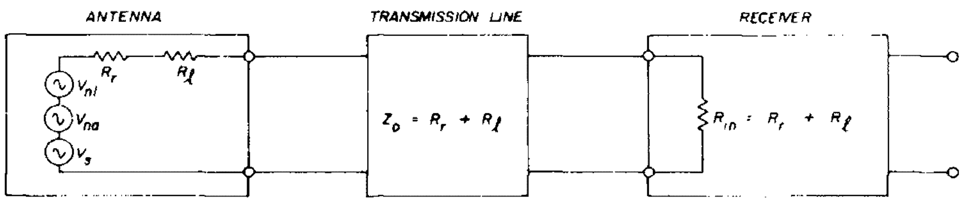



**fig. 3. Equivalent circuit of the basic receiving system. $V_{nl}$ is the antenna noise voltage, $V_{na}$ is the external noise voltage, $V_s$ is the signal voltage and $Z_o$ is the characteristic impedance of the line.**

magnetic wave and is related to the gain of the antenna by:

$$A = G_p \frac{\lambda^2}{4\pi} = \eta G_d \frac{\lambda^2}{4\pi} \qquad (5)$$

where $\lambda$ is the wavelength at the frequency of operation.

## the over-all system

With this background we are ready to redraw **fig. 1** into a complete schematic (**fig. 3**). Now we can determine the effect that antenna radiation efficiency has on receiver output SNR, and we can put some limits on this important characteristic.

Since a practical transmission line attenuates both signal and noise in equal proportions, transmission line loss has no effect on system output SNR as long as it is small enough that external noise dominates receiver noise. Remember that the important thing in the receiving sense is a high receiver output SNR. We can even

If you are interested in the derivation of **eq. 6** and **7**, send a self-addressed, stamped envelope plus 50 cents to cover Xerox costs to the author; a copy of the mathematical work will be sent to you. *editor*

In these equations $T_o$ is the reference temperature (room temperature, 270° Kelvin), $T_r$ is the receiver noise temperature, $T_a$ is the effective noise figure of external sources, and L is transmission line loss (L = 2 if line loss is 3 dB).

The receiver noise temperature is related to receiver noise figure by:

$$f_r = 1 + \left(\frac{T_r}{T_o}\right) \qquad (8)$$

where $f_r$ is the receiver noise *factor* ($f_r$ = 2 if noise figure is 3 dB).

The effective noise temperature of external sources, $T_a$, is related to an "antenna noise factor" by:

$$f_a = \frac{T_a}{T_o} \qquad (9)$$

where $f_a$ is the antenna noise *factor* ($f_a$ = 4 if antenna noise figure, $F_a$ is 6 dB). Data on $F_a$ are available for the entire world in reference 3.

**Eq. 6** and **7** give the *minimum* required for less than 1 dB over-all loss in output signal-to-noise ratio. Increasing beyond

this value provides absolutely no more than a 1-dB increase in SNR. In **fig. 4** the minimum efficiency defined by **Eq. 6** and **7** is plotted against antenna noise figure for representative values of transmission line loss and receiver noise figure.

We may have an antenna with a radiation pattern like antenna B in **fig. 2**—a perfectly satisfactory antenna for high-frequency use. In fact, not only is antenna B perfectly satisfactory, it will provide almost exactly the same system output SNR as antenna A, the efficient (and larger) antenna.

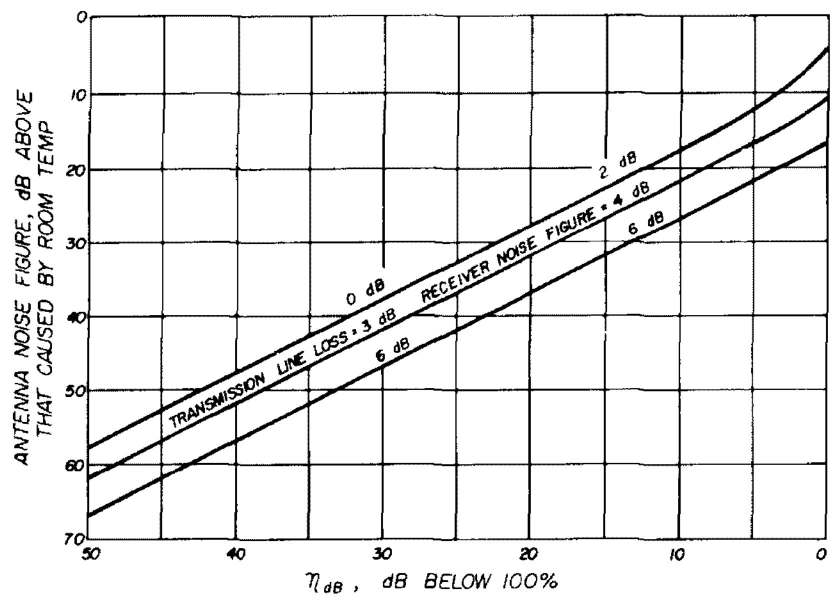



**fig. 4. Required minimum receiving antenna efficiency required for 1 dB degradation of system output signal-to-noise ratio, plotted as a function of antenna noise figure.**

Unfortunately, the three essential antenna characteristics—size, radiation efficiency and bandwidth—are mutually conflicting. That is, you cannot build an antenna with simultaneous high efficiency, small size and wide bandwidth. If two of these characteristics are essential to a given design, the third must be sacrificed. For example, if you want a small wideband antenna, then it must be relatively inefficient. If both efficiency and bandwidth can be sacrificed then the size of the antenna can be very small.

If external noise is inherently much greater than receiver noise (as it is in the high-frequency portion of the spectrum and below) antenna efficiency can be traded for small size and wide bandwidth. If wide bandwidth is not important, as in single-band amateur use, then the size can be reduced still further, and extremely small antennas can be practical. It is important to note here that we have not compromised directivity. Since we have only reduced radiation efficiency, the directive gain is unaffected even though the power gain may be less than unity.

Antenna B in **fig. 2** shows how increasing directive gain increases output SNR. Compared to an isotropic antenna, antenna B receives an identical amount of signal power, but off the back and sides where only noise exists it is less sensitive. As the back and side sensitivity is reduced further to increase directivity, SNR increases in direct proportion because of decreased received noise. Remember, if the system is external-noise limited, antenna directive gain governs output SNR, not power gain.

## antenna noise figure

The antenna noise figure, $F_a$, is a measure of all external noise sources combined: atmospheric, cosmic and man-made. When designing a high-frequency antenna, you have to look at the magnitude of the external noise expressed by $F_a$ to determine the required antenna efficiency. In fact, what must be found is the *minimum* expected antenna noise figure, so the realistic minimum required $\eta$

can be calculated. The minimum value of $F_a$ for the frequency range from 1 to 100 MHz is shown in **fig. 5**. Actually this is the cosmic-noise floor; noise figure less than the values shown in **fig. 5** will seldom be observed.

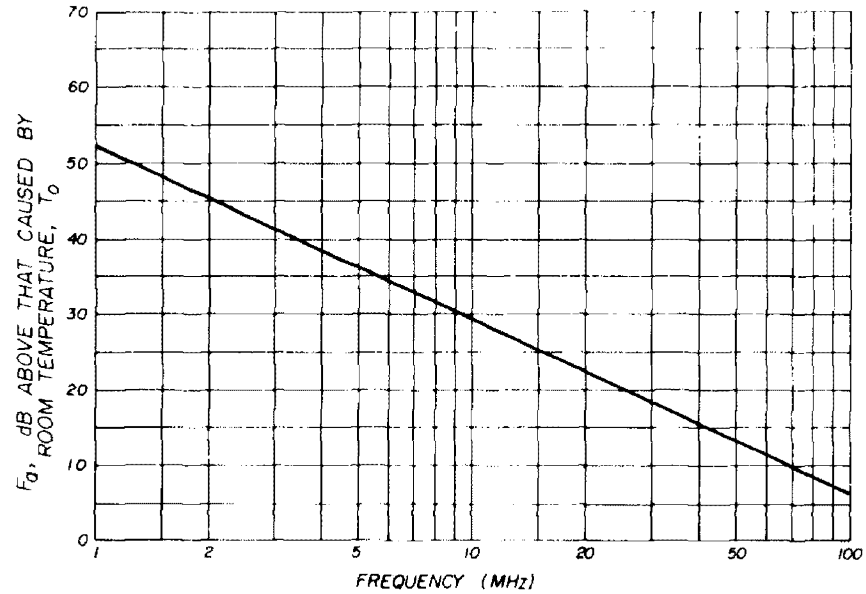



**fig. 5. Minimum antenna noise figure, due to galactic noise, as a function of frequency.**

## minimum required efficiency

If we combine **fig. 4** and **5** we can plot the required minimum antenna radiation efficiency at any frequency; this is the value which we must use in our design. At 7 MHz for example, with 3-dB transmission-line loss and a 4-dB receiver noise figure, minimum efficiency is −21 dB. This means that the radiation resistance of the antenna can be less than 1% of the total antenna output (or input) resistance. Loss resistance, $R_l$, can exceed ratiation resistnace, $R_r$, by a factor of 100 to 1!

This is a long way from the usual amateur receiving antenna and offers a terrific amount of design freedom. On the 160-meter band (1.9 MHz) with negligible line loss and a 2-dB receiver noise figure, loss resistance can exceed radiation resistance by nearly 10,000 to 1!

As wavelength increases high radiation resistance becomes more and more difficult to achieve with small antennas and losses increase. However, with the increas-

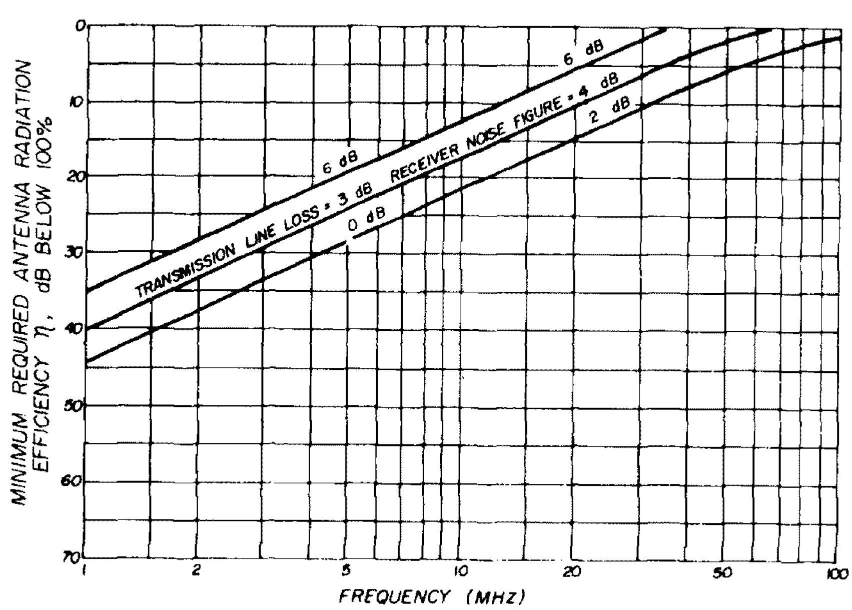



**fig. 6. Minimum required receiving antenna efficiency required for 1 dB degradation of system output signal-to-noise ratio, plotted as a function of frequency.**

ing external noise at lower frequencies, lower efficiencies can be tolerated, so radiation efficiency can be traded off for small size.

With smallness comes the ability to rotate the antenna. Thus we can build small directive receiving antennas that can be turned, even at 160 meters. This can make the difference between a marginal contact and armchair copy.

## local man-made noise

Man-made noise has ruined many contacts, especially on the high-frequency bands. Every active amateur who uses a vertical antenna on 40 meters can remember any number of contacts that were prematurely interrupted by automobile ignition noise or a neighbor's electric shaver. Let's examine the characteristics of man-made noise to see if there is a way to discriminate against it.

Man-made noise is characterized by predominantly vertical polarization, imbalance with respect to ground and an electric field form. Therefore, an unbalanced, vertically-polarized, electrical-field antenna is an inviting target— a quarter-wave vertical is ideal in this respect. On the other hand, an antenna which is horizontally polarized, magnetic-field sensitive and balanced with respect to ground discriminates against man-made noise but is no less sensitive to the desired far-field signal. Therefore it gives a higher system SNR.

It is not usually necessary for a receiving antenna to have all three of these desirable properties. If the antenna is balanced and magnetic-field sensitive, or balanced and horizontally polarized, it will usually discriminate against local noise to an acceptable degree.

## dipole antenna

The half-wave center-fed dipole can be miniaturized to provide a good receiving antenna simply by helically winding the two sides and adding a capacitive hat (fig. 7). This antenna, if mounted horizontally, will be horizontally polarized, electrical-field sensitive and balanced with respect to ground, and will discriminate rather well against local noise.

The over-all length of this antenna can be about six feet for all bands between 160 and 20 meters, although the number of turns on the helix will differ from band to band. Specific data for helical antenna design is given in reference 4. The antenna can be tuned with a grid-dipper or antennascope and fed with 72-ohm twin-lead.

If coax is used a balun should be used to preserve balance and electrical symmetry. The capacitive hats can be one to three feet in diameter and made of stiff wire; the helix can be one to two inches in diameter. This antenna will have the same radiation pattern as a full sized dipole (a figure 8 in the horizontal plane) so it has two nulls that can be used to reduce interfering signals.

If two of these antennas are spaced a short distance apart, say 0.05λ, and fed so that the phase difference between the

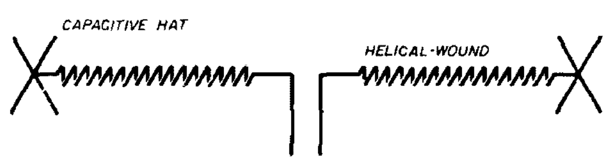



**fig. 7. Construction of a helical dipole, end loaded with capacitive hats.**

antennas plus the spacing equals a half wavelength or 180°, exciting things begin to happen. This is the basic end-fire array formula and results in a radiation pattern similar to antenna B in fig. 2. A 3-dB increase in SNR will result, and the front-to-back ratio will be quite good.

## the loop

Another antenna that can be used for receiving is the loop. A small loop antenna is primarily magnetic-field sensitive and balanced with respect to ground (if the plane of the loop is vertical the antenna is vertically polarized). For use on the 160-through 20-meter bands, a suitable loop can be from 2 to 3 feet in diameter; typical layouts are shown in fig. 8.

In fig. 8A the upper capacitor is used to resonate the loop; the resonant transformer is used to match the impedance of the loop to that of the transmission line. A match can also be obtained with the capacitive divider shown in fig. 8B. The loop itself can be made from one (or several) turns of wire or tubing. A good description of a loop antenna for receiving on 160 meters is given in reference 7.

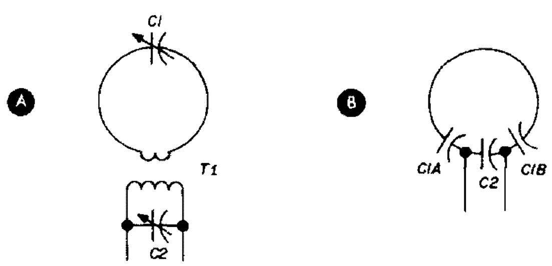



**fig. 8. Loop antennas. In A C1 resonates the loop; C2 tunes the matching transformer T1. In B C1 and C2 form a capacitive divider for matching to the feedline, and also resonate the loop.**

There is no reason why two or more loops cannot be phased the same as dipoles. The result will be a cardioid radiation pattern and an increased SNR. Loop spacing can be from 0.02 to 0.05λ, a little closer than dipoles; this is because the near field of a loop is more confined than that for a dipole.

Both the small dipole and small loop are fairly narrowband antennas so it is wise to use an antenna coupler between the transmission line and receiver. The antenna can then be peaked up at band center and the antenna coupler tuned for maximum response as the receiver is tuned across the band.

## summary

I have obtained good results with both a helical dipole and an untuned five-foot-long dipole with a capacitive hat when worked through an antenna coupler. However, the helical arrangement is somewhat better.

It is beyond amateur means to actually measure the radiation efficiency of an antenna, but there is a simple test which determines if an antenna has sufficiently high radiation efficiency to be a good receiving antenna. Simply hook the antenna up to the receiver; and if there is a significant increase in noise at the receiver output then the antenna efficiency is high enough.

At radio frequencies there is a complicated relationship between antenna size and radiation resistance. What it boils down to is this: if the greatest dimension of a linear antenna is greater than about four feet, and the area enclosed by a loop is greater than about two square feet, the result will be sufficiently high radiation efficiency.

It has been found through experience that when sky-wave circuits open up atmospheric noise levels increase to the point that antenna efficiency can be reduced another 20 dB at frequencies below 10 MHz. Since the sky-wave circuit is the rule for most amateur contacts the data of fig. 6 can be modified accordingly.

I hope this article has tickled your imagination enough to send you to the workshop, latest issue of *ham radio* in hand. Small antennas can do a superior *receiving* job on our high-frequency bands so give them a try.

## references


1. Robert L. Nelson, K6ZGQ, "A Primer of Basic Antenna Theory," *73*, May, 1968, p. 40.
2. James R. Kennedy, K6MIO, "The Real Meaning of Noise Figure," *ham radio,* March, 1969, p. 26.
3. "World Distribution and Characteristics of Atmospheric Noise," CCIR Report 322 (1964), International Telecommunications Union, Geneva Switzerland.
4. John J. Schultz, W2EEY, "Constructing Efficient Helical Antennas," *CQ,* May, 1968, p. 34.
5. Lewis G. McCoy, W1ICP, "The Army Loop in Ham Communication," *QST* March, 1968, p. 17.
6. "The Army Loop," Technical Correspondence, *QST*, May, 1968, p. 49.
7. Richard A. Genaille, W4UW, "A Top Band Loop Antenna," *CQ,* January, 1969, p. 41.


**ham radio**

# ground plow

This tool simplifies the installation of large radial systems that are required for efficient vertical antenna systems

Jim Ashe, W1EZT

**As I was working in the library** one day, researching a science article, I came upon a "ground plow." Now, what would you use a ground plow for? I wondered. I read.

The vertical antenna is a popular low-angle radiator; it is simple to build and very adaptable to multiband work. Lots of amateurs use vertical antennas, and I certainly would myself, if I had a place to put it. But if you want to build a really good vertical antenna, you wind up with a rather difficult grounding problem.

## some notes about grounding

An efficient vertical antenna *must* have an excellent ground system. Since it extends up into the air, there seems to be very little relation to the earth it rest upon, but actually the vertical antenna is *one-half of a dipole* **(fig. 1)**. *The other* half is a mirror image in the earth, made up of the electrical effects of heavy ground currents. But soil typically has a fairly high resistivity, and, outside tidal salt flats, the total ground resistance is too high for efficient grounding from a mere copper-clad steel ground rod. Much of the energy intended for the antenna is lost in the earth around it.

There is another way of looking at this **(fig. 2)**. Starting at the coax input cable, we find a complete circuit through the antenna's radiation resistance and its ground resistance. Current flows in loops around antenna circuits, just as in wire circuits, and in this case it flows from the cable center conductor, through the tower, into space, back to ground, and returns to the cable.

Some of the applied energy is radiated into space. This is accounted for in our diagram by the "radiation resistance," which can actually be measured as the antenna's radiated power divided by the square of the applied current. This is the familiar $W = I^2R$ equation. The radiated power can also be viewed as the product of current through, times voltage across, the radiation resistance. But this is not easily measured since the current that sees the radiation resistance also sees the

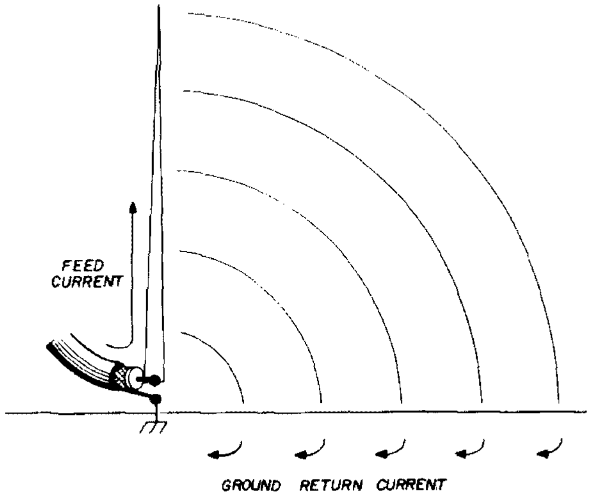



**fig. 1. Radio-frequency current flows into space from the tower and returns to the ground terminal.**

ground resistance and our measurements will indicate the total voltage across both resistances. (Fig. 2 supposes we are operating at resonance, where inductive and capacitive reactances balance out.)

Rf current flowing through the earth dissipates power in direct proportion to the ground reisitance. If ground resistance is equal to radiation resistance, one-half of the power reaching the antenna is lost. This applies to incoming signals, too. Since it is easier to make a lossy ground than a lossy antenna tower, this invisible part of the system is far more than a minor detail. If our antenna system is inefficient, we probably have to fix an inefficient ground.

It turns out that ground losses are really serious only near the tower. The losses decrease as the square of the current, which in turn falls off faster than inversely as the distance. At a part of a wavelength away from the vertical radiator, up to a few wavelengths at vhf, the current is low and the losses also low. We need a good ground surface only near the tower.

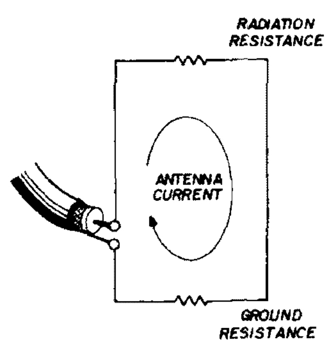



**fig. 2. Since the same current flows through both antenna and ground-return resistances, applied power is divided between the resistances. If the radiation and ground-return resistances are equal one-half the applied power is wasted as heat in the earth.**

A broadcast engineer's rule-of-thumb indicates how good this ground surface must be: "A radial ground system consisting of 120 quarter-wavelengths of copper wire will have an effective resistance of 2 ohms." Very few amateur ground systems will meet this spec, but by comparison one or two ground rods may be terribly inadequate.

Probably a great many amateur ground systems have resistances of 20 ohms or more; and that is in the radiation resistance ballpark for large vertical antennas. If the tower's electrical length is reduced the radiation resistance falls too as shown in fig. 3. This suggests a short antenna on 80 and 160 meters may have a radiation resistance in the one- to- ten-ohm ballpark! If you are thinking of serious communicating and particularly any DX work, it's very likely you need to put in some time on your ground system.

## the ground plow

Would you dig ditches to install a hundred and twenty radials four to twelve inches in the ground? I am afraid I wouldn't. At 80 meters that would be about 7200 feet, or twenty feet per day for a year. The wire is probably not much of a problem, and for this application I'd try the power company, junk yards, surplus dealers, and other such sources for good used copper wire. It needn't be all the same size but should be around number 10 or so for adequate strength.

The pros put down a good ground plane with a ground plow (fig. 4). Its sharpened leading edge digs down into the ground until the skids or wheels at the ends of the main frame are riding along the surface. A tractor pulls the plow, but it should not be so close that it yanks the plow out of the ground. One or two chimney blocks or some cinder blocks serve as weights, and you walk

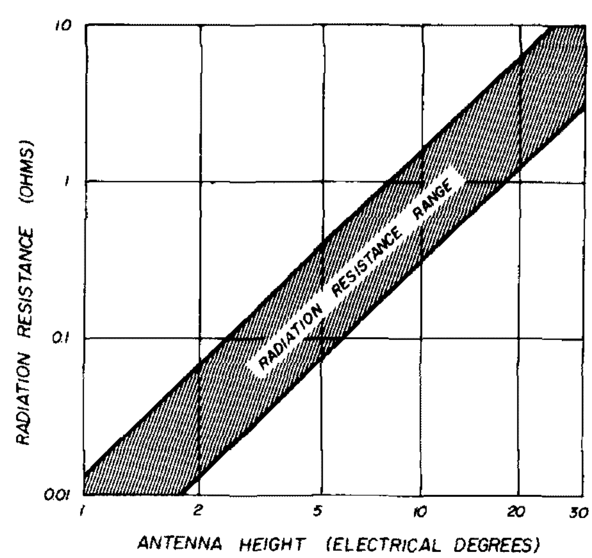



**fig. 3. Radiation resistance of short vertical antennas.**

along behind guiding it in a reasonably straight line but around rocks and obstacles. Your helper rides the tractor, which point, of course, is to make this a small group or a club project, rather than do it all yourself.

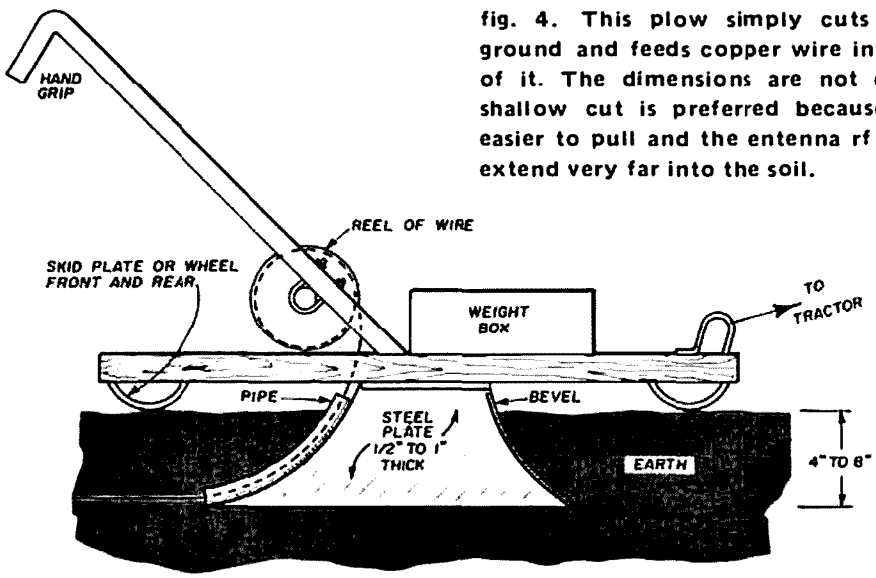



**fig. 4. This plow simply cuts a slit in the ground and feeds copper wire into the bottom of it. The dimensions are not critical, but a shallow cut is preferred because the plow is easier to pull and the entenna rf field does not extend very far into the soil.**

for a shallow cut in good soil could be a small garden variety. Extra weight may be required on the driving wheels. With this gear an 80-meter ground plane should go down in a half day.

How about making the plow? This is easy if there is a welding shop nearby. Since the plow will be used rarely it can be assembled from scrap iron. A cutting torch will cut inch thick iron like wax and can rough out the bevel too. Then the bevel is finished with a hand grinder and the rest of the frame, skids, and etc. welded on. The finished plow can be carried around in a station wagon. A key

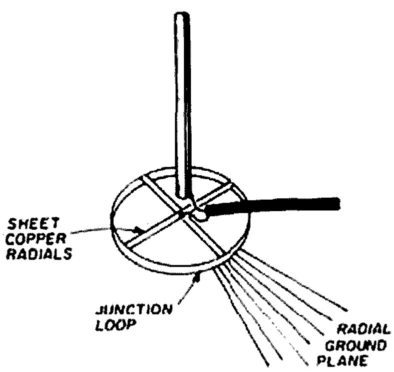



**fig. 5. A simple junction loop assembly with radials connected to a ring rather than to the tower. This provides a reliable structure and avoids problems with big bundles of heavy copper wire.**

## finishing up

Unless you are using a gamma match your tower will rest on an insulator in the center of the ground plane. Broadcast practice is to have a heavy expended-mesh screen in this area since the voltage and current levels are very high, but you can get by with some two- or three-foot sheet copper radials to carry the ground return currents.

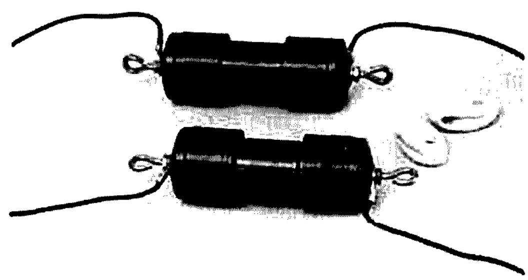
Outside the radials there is a continuous ring of copper pipe or more copper strips. Two to four copper-clad steel rods are driven into the earth around the ring's perimeter. These serve two purposes: they hold the ring in place and serve as an additional channel for carrying lightning strikes into the earth. Your antenna is likely to be hit by lightning at least once every year or two, and if you are in certain parts of the Midwest or Florida, you can expect frequent fireworks.

The radials are brought to the circular ring, rather than to the tower itself, and connected by soldering or brazing. This finishes the job, and you will have an effective ground system that will enhance communications.

ham radio

# new products

## multi-band antenna

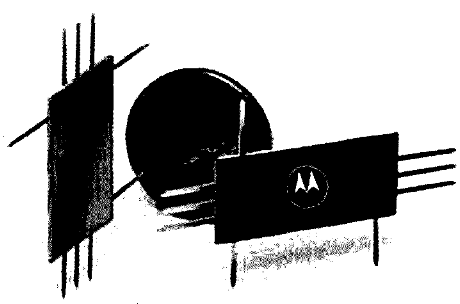


The Bomerang II antenna system is designed for 80 and 40 meters, but when tuned properly, it will also perform satisfactorily on 20, 15 and 10 meters. The antenna system consists of 108 feet of copper wire, two high-strength lightweight traps, a 1:1 balun, hardware and insulators. All hardware is stainless steel or copper and the traps are encased in impervious polyvinylchloride.

This antenna is usually erected in the popular inverted-V configuration, but sometimes it is installed as a flat-top. Because of the low weight of the system and the low wind profile of the traps, number 16 copper wire may be safely used; number 12 or 14 may be used for high strength. Suitable number 16 wire, such as Belden 8074, (126 feet) costs about $3.00.

The Boomerang II antenna system, rated at 2 kW PEP ssb, complete with balun, less wire, is $22.95. Boomerang II without balun is $12.95; the AB1 Boomerang 1:1 balun is $10.95. All prices postpaid in the U. S. A.

## ic duplexer



The new Motorola integrated-circuit duplexer, the MCH5890, operates at frequencies between 400 and 500 MHz with up to 40 watts input and features 0.1 dB transmit-mode insertion loss with a typical 25-dB transmit-mode isolation figure. Although the primary job of the MCH5890 is as a transmit-receive switch, it will also find use as a monitor network in transmitter circuits, as the sampling unit in afc and agc circuits and other related communications applications. For more information, write to Technical Information Center, Motorola Semiconductor Products, Inc., Box 20924, Phoenix, Arizona 85036.

## portable ssb transceiver

One of the hottest new items at the SAROC convention in Las Vegas in February was the new TR-5 ssb transceiver shown by the R.L. Drake Company. This new transceiver, planned for 1970 deliveries, and targeted for the $400 price class, features five-band coverage, nearly all

solid-state construction (the three vacuum tubes in the unit are for the driver and power amplifier) and a built-in power supply that can operate from 12 Vdc or 120/240 Vac. New pulse-power techniques give 500 watts PEP input on ssb and 100 watts on CW while running less than 2 watts input during tune up.

Frequency stability is less than 200 Hz after warmup. Dial calibration—with concentric dials—is zero to 1 MHz in 1 kHz divisions. Receiver sensitivity is 0.5 μV for 10 dB signal- plus noise-to-noise ratio. Receiver agc results in a 3-dB change in output for 60-dB change in signal level. A built in meter reads plate current and S units. Transmit modes are upper sideband on 10, 15 and 20 meters, lower sideband on 40 and 80 meters, push-to-talk only (manual cw). Accessories include an i-f type noise blanker and a 100 kHz calibrator. R. L. Drake Company, 540 Richard Street, Miamisburg, Ohio 45342.

## fast-sweep knob

A new fast-sweep control knob that dials like a telephone has been introduced by Kurz-Kasch, Inc. The new knob features a recessed finger pivot on its perimeter that eliminates the awkward crank handle of earlier designs. The new fast-sweep knob measures 2½ inches in diameter and 1 inch deep, and is molded from black plastic with a brushed aluminum inlay. Price is $3 each form larger electronic distributors.

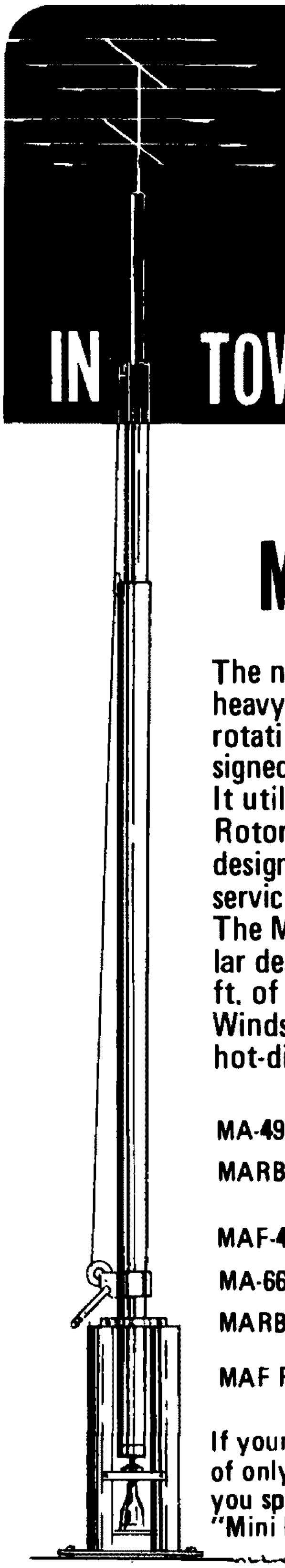
### coil calculator

The American Radio Relay League has introduced a modern slide-rule type calculator that can be used for solving problems dealing with inductance, capacitance and frequency. If you remember the old Type A calculator, this is an up-to-date easier-to-use version. The L/C/F calculator efficiently solves simple problems such as the frequency corresponding to a given wavelength, and vice versa, as well as much more complex problems such as determining inductance and capacitance values that will resonate at a given frequency (or wavelength), finding the resonant frequency for a given combination of inductance and capacitance, calculating the size coil required for a given inductance when coil length and diameter are predetermined, and finding turns-per-inch for various wire sizes. Scales cover inductance for 0.1 to 1500 microhenrys capacitance from 3 pF to 0.01 $\mu$F and frequency from 300 kHz to 100 MHz.

The L/C/F Calculator, Type A, a convenient bench-top 4 x 10 inches, is $2 from your local electronics store, or direct from the American Radio Relay League, 225 Main Street, Newington, Connecticut 06111.

### electronic circuit design handbook

The new third edition of this well known handbook contains well over 600 proven and tested electronic circuits—circuits selected by the editors of *EEE Magazine* on the basis of their originality and practical application. Circuits represent all branches of electronics, and include both basic and advanced designs that can be used to suit the reader's specific needs. Included in this handbook are control circuits, regulator circuits, filter and suppression circuits, amplifiers and oscillators, converters and inverters, power supplies, gating and logic circuits, as well as many others. 384 pages. 750 illustrations. $17.95 hardbound from Tab Books, Blue Ridge Summit, Pennsylvania 17214.



*only* in repeater systems with exactly 5-MHz channel spacing.

Other paragraphs in the docket concern attendance at the repeater transmitter or at an authorized *fixed* control point, and limiting power input (to repeaters) to 600 watts.

Although the proposed rules have the effrontery to suggest that they won't duly inhibit the growth of useful repeater systems, how can a system possibly grow when all of the parameters are initially specified? The amateur regulations have traditionally been more permissive than restrictive—a policy that has bred initiative and inovation. Development of interlinked repeaters, multiband systems, tv and wideband data repeaters, command-control systems and telemetry will be stifled by the rules proposed in Docket 18803, as will experimentation, one of the mainstays of amateur radio.

There's still time to do something about Docket 18803, but you've got to stand up and be counted *now*. All of our vhf fm readers should already have received a copy of the docket which went out several weeks ago, along with the comments of our fm editor, Jay O'Brien, W6DGO, and a resume of related ideas generated at a recent meeting of the California Amateur Relay Council. If you haven't seen a copy of this material, but would like one, simply send me a stamped self-addressed envelope and I'll put a copy in the return mail.

Comments on Docket 18803 must be filed by May 15, 1970, with reply comments on or before June 1, 1970, so you don't have time to dilly-dally. Remember that comments to FCC proposals must include an original and fourteen copies (complete filing details are included with the Docket 18803 mailing piece from *ham radio*). With the abundance of office copiers that one finds today, those 14 copies shouldn't pose too much difficulty to the amateur who is really interested in contributing his ideas.

Jim Fisk, W1DTY
editor

75 cents

hr

focus on communications technology . . .

# ham radio

magazine

JUNE 1970

## communications experiments with light-emitting diodes

*this month*

june 1970
volume 3, number 6

# ham radio magazine

# contents



## staff



## offices

Greenville, New Hampshire 03048
Telephone: 603-878-1441

**ham radio** magazine is
published monthly by
Communications Technology Inc.
Greenville, New Hampshire 03048

Subscription rates, world wide:
one year, $6.00, three years, $12.00.
Second class postage
paid at Greenville, N. H. 03048
and at additional mailing offices

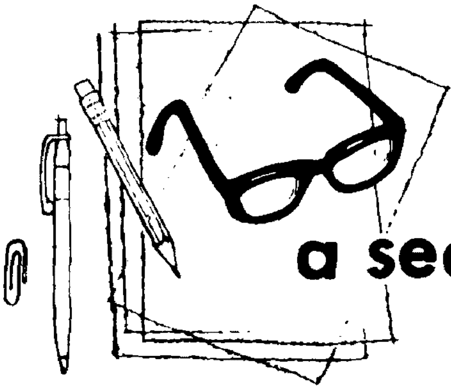
Foreign subscription agents:
United Kingdom:
Radio Society of Great Britain,
35 Doughty Street, London WC1, England.

All European countries:
Eskil Persson, SM5CJP, Frotunagrand 1,
19400 Upplands Vasby, Sweden.


Printed by Wellesley Press, Inc.
Wellesley, Massachusetts 02181, U.S.A.

**ham radio** is available to the blind
and physically handicapped on magnetic tape
from Science for the Blind,
221 Rock Hill Road, Bala Cynwyd,
Pennsylvania 19440.
Microfilm copies of current
and back issues are available
from University Microfilms,
Ann Arbor, Michigan 48103.

Postmaster: Please send form 3579 to
**ham radio** magazine, Greenville,
New Hampshire
03048



# a second look

by jim fisk

**Perhaps you've noticed** the articles in the national news media aimed at educating the lay public about the electronic computer and its impact on modern society. Some of these news items are well done and reflect the responsibility of the reporter. Others, while leaving something to be desired in authenticity, nevertheless provide some amusing reading. An example of the former is the extensive coverage given to the Apollo 12 lunar landing module navigational computer and its vital role in the success of the mission. These reports did a pretty good job of informing the public about a complex subject in a nontechnical manner. On the other hand, articles in the same publication described the computer as some kind of electronic monster bent on invading personal privacy and generally disrupting the status quo of the credit consumer and taxpayer (i.e., you and me).

Regardless of what the public is led to believe in the national news media, the fact remains that the computer and the vast industry behind it are a solid part of the American scene and will be around for awhile.

A casual glance at the electronic industry trade journals will give an idea of how rapidly computer technology is changing. The intense competition for new markets has resulted in innovations, new devices, higher speeds, and more efficient circuits.

An area of computer innovation not generally known to the public is interactive computer graphics. Simply stated, the technique consists of using hardware designed to enable the engineer to communicate directly with the computer. In a typical application, the engineer watches the problem on a crt screen as the solution is being developed. He can communicate directly with the computer through a light pen and a typewriter. The typewriter allows him to carry on a conversation directly with the computer; he uses it to respond to cues displayed on the crt screen. The light pen, which is a pencil-shaped bundle of fiber optics coupled to the crt, allows the engineer to precisely place lines or circles of desired dimensions on the crt screen, create new shapes, or change the display to the desired scale—all in real time.

The beauty of interactive computer graphics lies in the phrase, "in real time." Using former techniques, the engineer was required to design a program, run it, study the printout, then iterate changes through the system until an acceptable solution was obtained. At upwards of $600 an hour for computer time, it's easy to understand why this new innovation of computer technology has been received with such enthusiasm.

**Jim Fisk, W1DTY**
**editor**

# gallium arsenide LED experiments

Theory and application of light-emitting diodes and photodetectors in communications circuits

Ralph W. Campbell, W4KAE, 316 Mariemont Drive, Lexington, Kentucky 40505

One of the most interesting fields of electronics that has been neglected by amateurs is communications by lasers and their relatives, the light-emitting diodes. Experimental data is meager in the amateur literature. Ready-made equipment, of course, doesn't exist.

I became interested in building a diode laser communications link between my home and a local broadcast station, both for the novelty of the idea and to communicate with colleagues. The experiments reported in this article are a first iteration toward achieving the laser link.

The professional literature describes bulky, expensive optical devices as modulators placed in front of equally expensive ruby rods or gas lasers. The light-emitting diode (LED) and diode injection

**Broad-area quadrature detector for broadcast-band response. A 1-MHz pilot carrier, injected through the lens shade, produced sidebands above and below 1 MHz in a Hallicrafters SX-122.**

laser are well within amateur means, so I investigated those.

My early experiments were made with equipment designed from scratch and with a minimum of help from reference material, because I wanted to learn by doing. After much trial and error, I was able to achieve what I believe is an amateur world record for one-way ranging with LED's: a distance of 250 feet. This was done with the equipment described in the following paragraphs. I used a GE SSL-4 LED transmitter and a broad-area detector comprised of seven GE L14A502 phototransistors in a honeycomb matrix.* The 250-foot range is five times as far as that obtained with commercially designed equipment using a white (incandescent) noncoherent light source and the same type detector.

*The SSL-4 is available from G. E. for about $7.50. Write Miniature Lamp Dept., General Electric Company, Nela Park, Cleveland, Ohio 44112. The L14A502's were obtained from G. E. as a sample

†The gunsight ($10.00) and achromat (about $3.50) are available from Edmund Scientific Company, 101 E. Gloucester Pike, Barrington, New Jersey 08007. The CV-148 ($5.00) can be obtained from John Meshna, Jr., Box 62, E. Lynn, Massachusetts 01904.

## a word of caution

Unless you've had experience with laser diodes, I'd recommend starting first with LED's. An early attempt with injection lasers ended in disaster. These devices must be operated with a power supply duty cycle of 0.1 percent—preferably less. The average power of a diode laser must be kept low to avoid overheating, so a pulsed power supply is used. If its duty cycle is much above, say, 0.05 percent the injection laser will go up in smoke no matter how good its heat sink. Cryogenic cooling allows higher duty-cycle pulsing, but this is beyond the means of most hams.

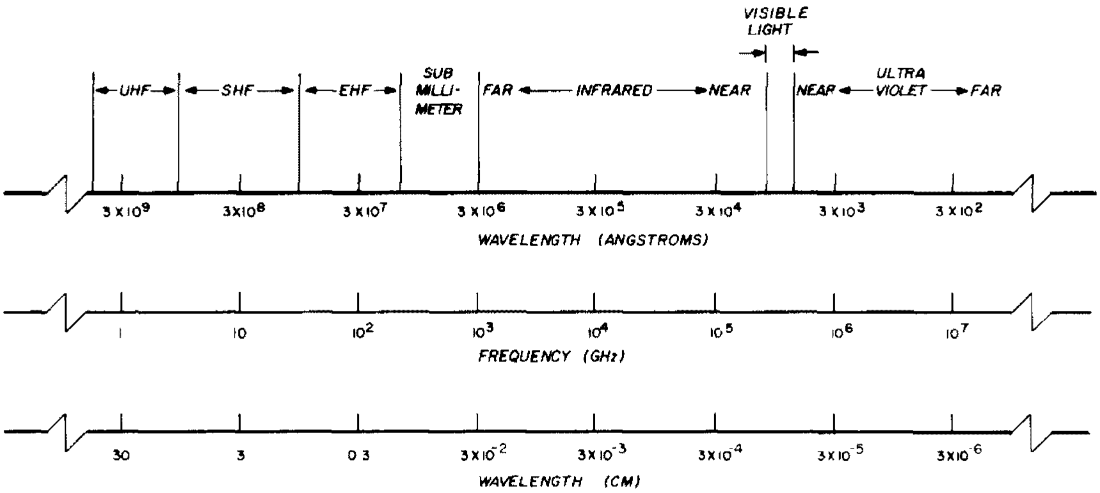



**fig. 1. Approximate spectral regions. "Near" IR and UV spectra are referenced to visible light.**

## operating frequency

The equipment used in these experiments operated in the infrared region (**fig. 1**). Because of the nearness of infrared to the visible light portion of the spectrum, optical equipment can be used to enhance operation. The optics I used were a surplus NC-3 gunsight, an f/1.9 two-inch-diameter achromatic lens, and a surplus British type CV-148 sniperscope image tube.† The gunsight was used to collimate light rays from the LED; the achromat was used to focus light input and output.

The image tube was used as an infrared detector in some experiments. Its sensitivity is low for LED work, but it's a good detector for laser use.

## how LED's work

When driven by an external power source, certain semiconductor materials produce light. The light can be in the visible or invisible part of the spectrum, depending on certain laws of quantum mechanics. If an LED's p-n junction is forward biased, electrons from the n region (valence band) will flow across the energy-band gap into the p region (conduction band). Here the electrons recombine with electron holes, then fall to a lower energy level where they emit a photon, which is a quantum of light.

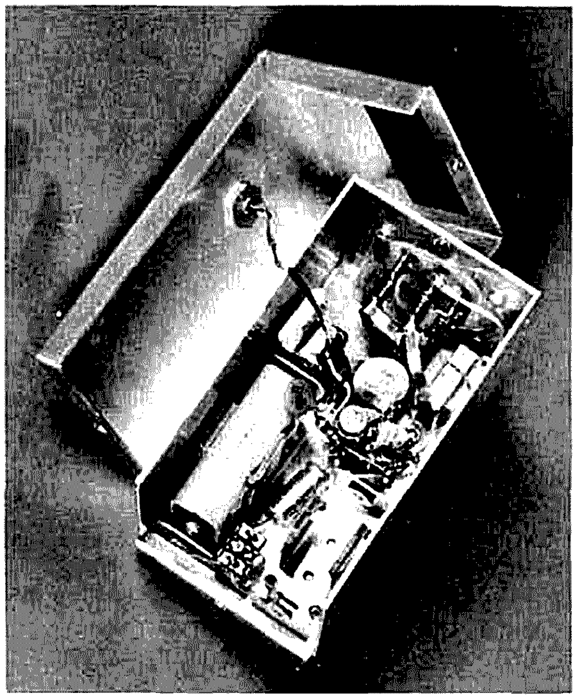
The wavelength of the emitted photon depends on the energy-band gap between the n and p regions. LED's constructed of gallium arsenide (GaAs), as used in my tests, have a band gap that permits radiation in the near infrared region (referenced to visible light; fig. 1). Visible light may be emitted from LED's with a wider band gap. For example, LED's made of gallium phosphide emit green light.

LED output is determined by the geometry of the host material pellet and the device's packaging. Some GaAs LED's are packaged with a parabolic reflector as part of their structure. An epoxy lens collimates the light output to a very narrow region of the device's optical axis.

LED light output is noncoherent, whereas that from a laser is coherent. The significance of these terms will become apparent when we consider the laser, discussed next.

## lasers

The first operating laser was demonstrated by T.H. Maiman in 1960. Its principles have been covered extensively in the literature. The following brief description, although considerably simplified, is given to show the comparisons



**Improved Broad-area detector using seven GE L14A502 photo transistors connected as diodes. A maximum range of 250 feet was obtained at twilight with this unit. A surplus NC-3 gunsight was used as a collimator.**

### glossary

| | |
|---|---|
| responsivity | output signal per unit input signal of photodectector. |
| sensivity | change in output per unit input change of a photodetector. |
| spatial coherence | phase relationship between two wave trains in space (determines output directivity of a laser). |
| spectral coherence | a measure of the restriction of a photodetector's light output to a single wavelength or band of wavelengths (i.e., color response). |
| lasing | phenomena exhibited by certain materials when the threshold condition has been achieved for self-sustaining photon emission. |
| monochromaticity | degree of response to one color in the electromagnetic spectrum. |

between LED's and lasers.

Laser is an acronym for "light amplification by stimulated emission of radiation." In its most prevalent form, the laser is used as an oscillator. It can be used as an oscillator. It can be used as an amplifier, but its spontaneous (noncoherent) emission is so great that it doesn't perform well as an amplifier at low input levels.*

An LED, as explained above, emits light due to the transition of electrons between energy levels. However, the light output is noncoherent, which means its phase and amplitude are not correlated (recall that if two wavetrains of the same frequency are in phase, their amplitudes are maximum).

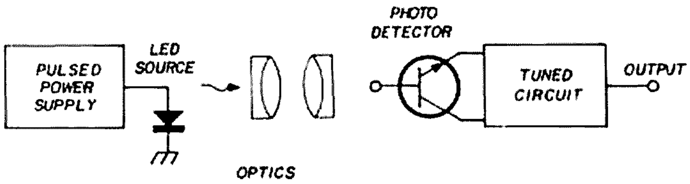



**fig. 2. Basic arrangement for a simple pulsed light system. Achromat lenses increase performance, but are difficult to align if separation between source and detector is great.**

In a laser, radiation from an external source, called a pump, raises the electrons of the active material from a lower to a higher energy level. Some of the electrons are absorbed, but some will be driven downward to an intermediate level, from which they will return to their original low-energy state and emit photons. This process, called stimulated emission, continues in a chain reaction similar to a nuclear explosion. If the laser output is coupled to a resonant circuit of high Q, and if certain optical arrangements are used, an extremely powerful and coherent light source is produced.

The difference between laser and LED operation is that LED's do not exhibit stimulated emission, and their light output is spatially noncoherent.

*Thus the term "loser" is sometimes used; i. e., "light oscillation by stimulated emission."

## coherence

Two types of coherence are involved: spectral and spatial. The former determines how closely light output is restricted to a single wavelength or band of wavelengths, while the latter pertains to waves in space. Spectral coherence is a measure of monochromaticity. Most LED's and all lasers produce monochromatic (single-color) light. Since spatial coherence depends of the frequency of pumping radiation, which does not occur in LED operation, LED's exhibit little spatial coherence.

If the input wave to a laser (supplied by the pump) has plane wave fronts, the laser's output will also contain plane waves. The degree of phase correlation (or regularity) between these plane wave fronts is a measure of spatial coherence, which determines directivity. The laser beam will have almost constant width for a distance, S, according to the following relationship

$$S = \frac{D^2}{4}$$

where D is the diameter of the laser output source, and $\lambda$ is the wavelength of radiation. Beyond this distance, S, the

**The SSL-4 LED mounted in an Amphenol MX-1025/U cable termination, encapsulated with epoxy (right). The surplus CV-148 image tube is shown at left; the battery supplies 510 volts for its photocathode.**

wave begins to assume a conical shape with wave fronts assuming a sphere, as in radiation at radio frequencies.

## pulsed light systems

A simple pulsed light system using LED's is shown in **fig. 2**. This basic system can be modified to include a speech modulator and various postdetection amplifiers, as well as the optics shown focus divergent light to allow wider spatial separation between LED and detector. The theoretical separation is of the order of several miles. In practice, this is limited to several yards because of the difficulty in aligning the optics and LED response to stray fields.

## detectors

The limiting element in an LED communications system is the detector. The best detectors for light transmission have high responsivity and sensitivity. Minority carrier lifetime determines responsivity, whether we're discussing photodiodes, photo transistors, or image tubes. A detector with high responsivity will deliver an output signal within picoseconds after

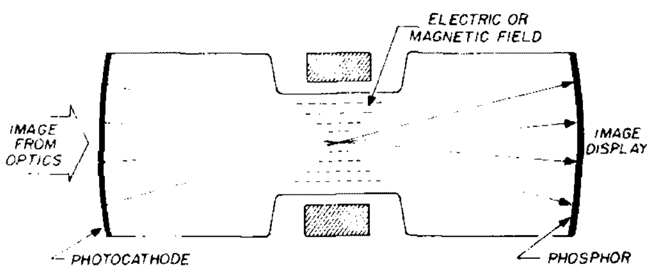



**fig. 3. A typical electron image converter tube. These respond well to pulsed light sources but sensitivity is low for LED work.**

an input signal is applied. Sensitivity is a measure of the change in a detector's output per unit input signal change. Some authorities use the word "responsitivity" to describe detector performance—a coined word combining both terms.

## broad-area response

Another important consideration for light detectors is broad-area response. A simple analogy using the human eye will explain this. If you've ever been on a hilltop at dusk you may have noticed how bright an incandescent lightsource appears on the skyline. Here we have many point sources of light that are partially

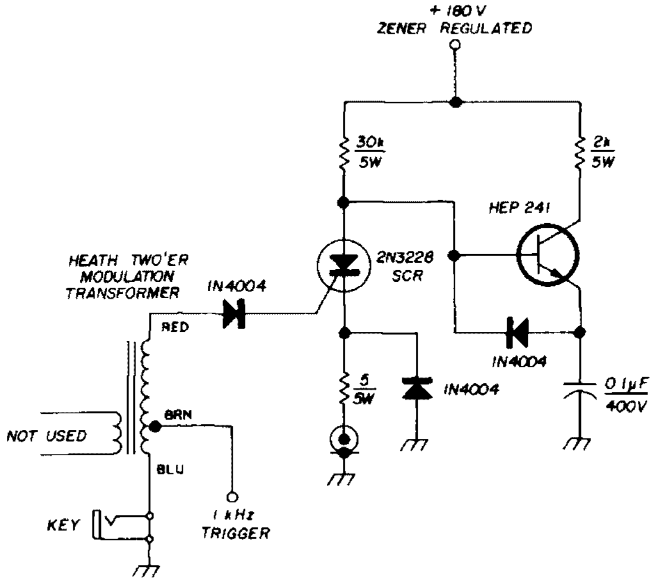



**fig. 4. A pulsed power supply for LED's that develops about 900 mW. It's not suitable for an injection laser, however.**

collimated into parallel rays. With sufficient sensitivity, there's almost no limit to the amount of incident light falling on the broad-area retina of the eye. A photo detector should have an array of at least ten input sources to receive noncoherent radiation from an LED.

I made my broad-area detector array by arranging several GE type L14A502 photo transistors in a honeycomb matrix. These were connected as photodiodes in a lens-in-can arrangement to provide optical gain.

## detector response

Photo transistors (or more correctly, photo duo-diodes) are of the npn type. When operated in the reverse-bias mode, I consider them to be p-n diodes. They respond to rms output only. Peak power, as from a pulsed system, can't be sensed by these devices, even with a 6 percent duty cycle. The only diodes that respond to peak input are the p-i-n types such as the

Schottky barrier diodes. However, these are rather expensive.*

The cadmium selenide (CdSe) photo cell has relatively high sensivity, but suffers from the same disadvantage as the common npn photo duo-diodes: linear response. An envelope detector must be used with these for best results. Cadmium selenide cells exhibit a spectral shift further into the visible light region. You must, therefore, make a tradeoff between higher sensitivity and spectral response when considering the CdSe cells.

## the image converter tube

The S1 image tube is a good candidate for an infrared detector. A simplified sketch is shown in fig. 3. These tubes respond to pulsed light because the phosphor screen (made of Willemite P1 phosphor) has a storage effect on the recombined secondary carriers received from the photocathode. The storage phonomenon results from a sort of mathematical integration of impinging elec-

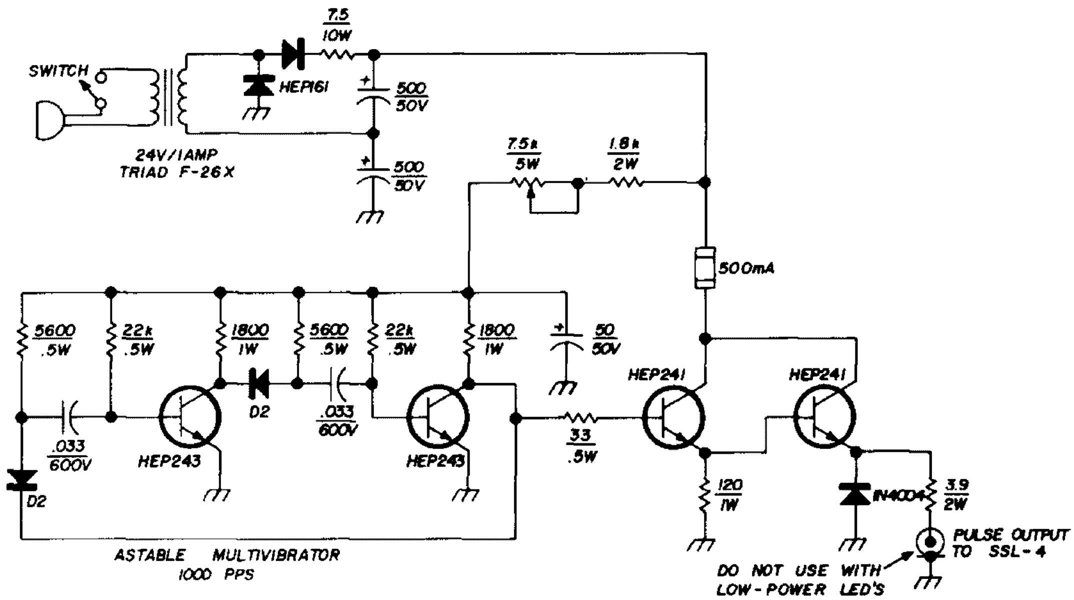
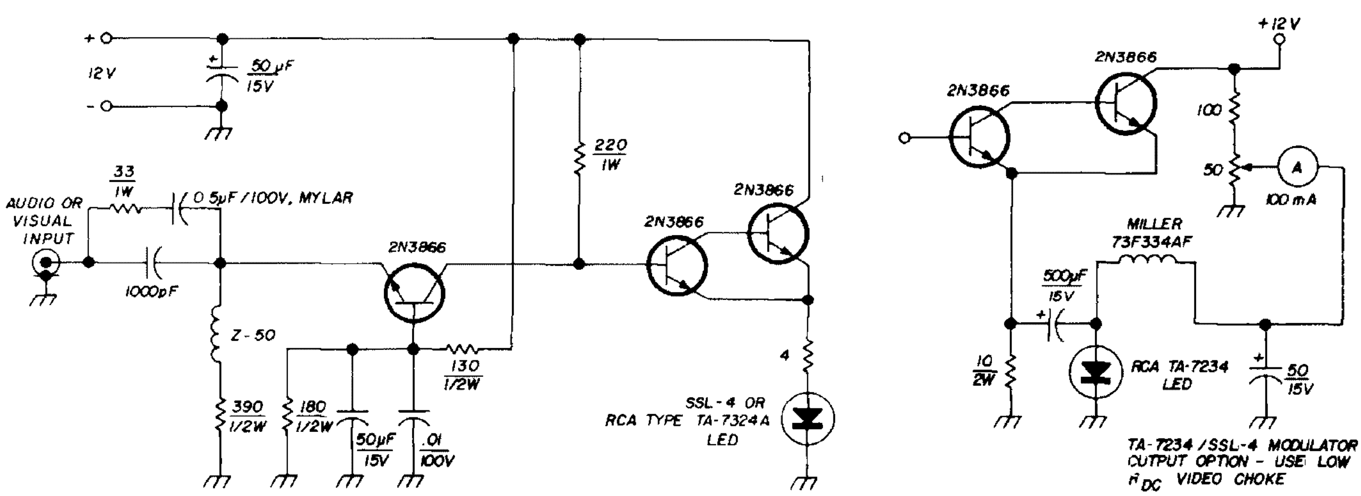



**fig. 5. A power supply good for 1.25 watts. Duty cycle is 50 percent. Speed-up diodes D1 and D2 shape the output pulse trailing edges.**

*The type PIN-5 ($24.00 post-paid from United Detector Technology) seems to have greatest responsivity for pulsed light.

**fig. 6. An LED modulator. Circuit on right can be used for improved voice modulation (per JA3HA).**



trons. Image tubes that respond to repetition rates of the order of 100 pps are made of P20 phosphor. Nonimaging photodiodes, properly biased, operate similarly on pulsed light: the Schottky barrier is an equivalent mechanism to the image tube phosphor in its storage effect.

## power supplies

A pulsed power supply that will deliver 900 mW is shown in **fig. 4**. Its duty cycle is about 1 percent. When the capacitor is fully charged the HEP-241 bipolar switches the pulses through the scr, which acts as a gate. The result is a current pulse less than 200 nsec. Diodes D1 and D2 shape the trailing edges of the pulses.

The series resistor in the output should be 4, 6 or 10 ohms when the supply is used with LED's such as the ME-2, SSL-4 or TA-7324A respectively. I'd recommend using a no. 248 miniature lamp to replace the LED when testing a new light-source design. It's a lot better to blow up a lamp than an ME-2 at $30.00 each.

## modulators

The modulator shown in **fig. 6** is patterned after a circuit in G.E. Solid-State

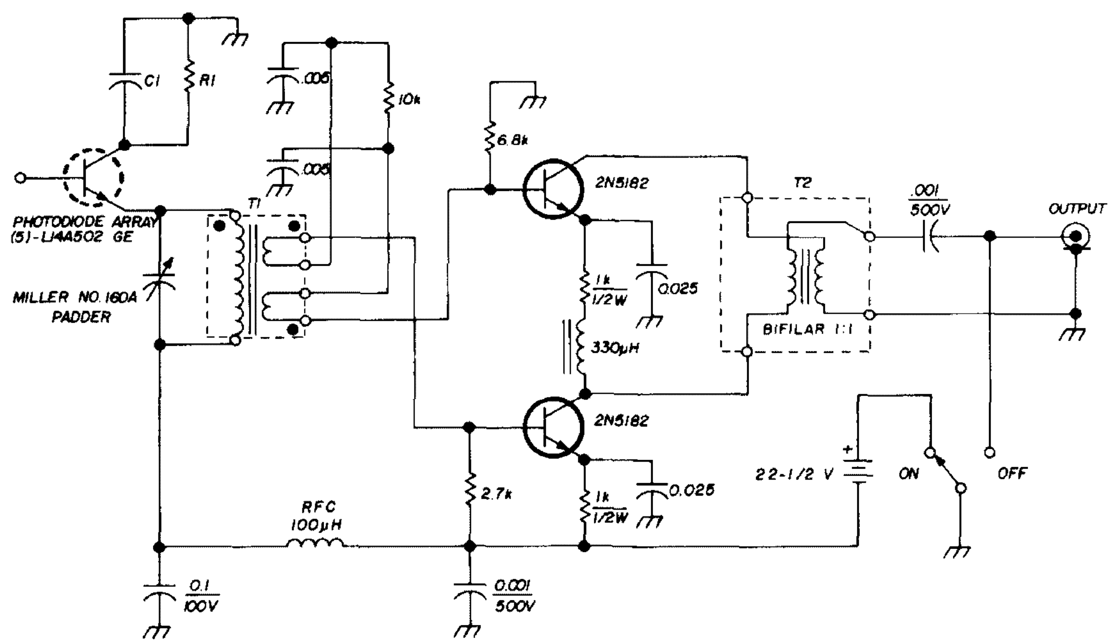



**fig. 7. Quadrature detector used for receiving LED output translated to the broadcast band. C1 is five .005 capacitors in parallel; R1 is five 560 ohm resistors in parallel. T1 is a tuned quadrature coil wound on 57-1736 carbonyl sf toroid; primary is no. 32 wire wound to fill 7/8 of the core; secondaries are 1/8 of the core's circumference. T2 is a 1:1 quadrature coil and consists of no. 32 wire wound to fill a 57-1736 toroid.**

whose amplitude is determined by dynamic response of the thyristor. Pulse width is determined by the supply's dynamic resistance and the 2k resistor. This circuit is very good for 70-ampere-peak, high-threshold emitting devices but is unsuitable for injection lasers.

A 50 percent duty-cycle pulsed supply is shown in **fig. 5**. The heart of the circuit is an astable multivibrator. The supply puts out square pulses with a rise time of Lamp Manual No. 8270. I made changes to allow use of npn overlay transistors in place of the original pnp's. This circuit will work between audio and video frequencies. Shielding may be a problem when using the modulator with LED's.

The modification shown on the right of **fig. 6** is based on experiments by JA3HA for voice modulation. The LED on the right is forward biased to half maximum cw current value, with no modulat-

ing signal present. A 500 Hz tone is then applied, and average forward current is increased to 2 amps. (The LED *must* be in a heat sink.) The 500-Hz source is then replaced with a speech amplifier. Voice peaks will not cause lasing in the LED.

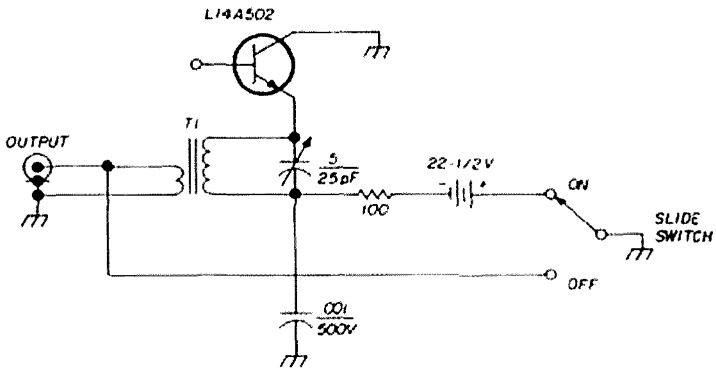



**fig. 8. Simple photodetector for 50-MHz response. One or more photo transistors can be used, depending on the tuning method. The primary of T1 is 2 turns, secondary is 7 turns on a Permacor 57-6075 toroid.**

## quadrature detector

A successful detector for output in the broadcast band is shown if **fig. 7**. The circuit is similar to a 1:1 balun, except that output is obtained at two ports from a single-port input. All three ports, if used, have a common ground. Ninety-degree phasing exists between the two output ports, netting 180° as in a conventional balun. Advantages are reasonably high gain, no neutralization, and limited bandwidth at the lower resonant frequencies. It also has low noise response.

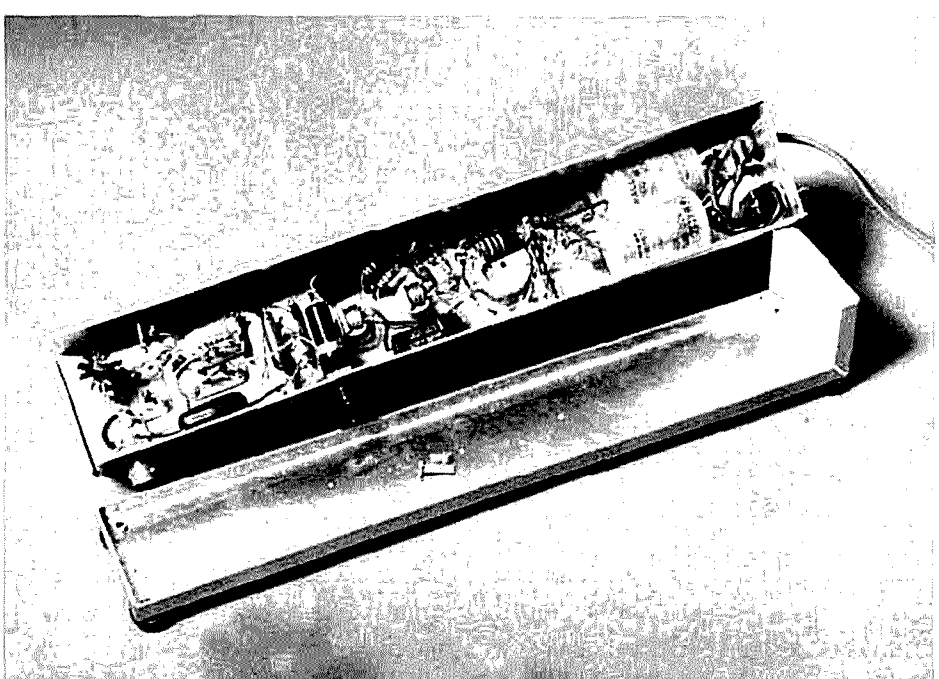
The optical input to this detector consisted of five G.E. L14A502's operating nonlinearily (rms response), connected in a matrix.

## detector for 50 MHz

A simple circuit for one or two photodiodes is shown in **fig. 8**. Parallel instead of series tuning is suitable for one photodiode. With a junction capacitance of 2 pF, it should be possible to tune as many as five photodiodes in parallel. In a series-tuned arrangement, 10 pF should allow response to perhaps 400 MHz.

## broad-area detector

My most successful broad-area detector used seven GE L14A502 photo transistors connected as photodiodes (**fig. 9**). Component values were chosen to bring the response of the IC into the audio region. The surplus tape head shown in



**The 1.25-watt LED power supply. This supply runs quite hot, and should have a 2.5-ampere transformer and 2.5-ampere epoxy rectifiers. Wakefield 254SI heat sinks are used for the astable multivibrator transistors.**

the schematic couples output into high-impedance headphones.

## conclusions

It is difficult to pulse modulate LED light beams. Audio and video modulation is practical at 1000 feet or less; however, the limiting parameter is stray fields. Another factor is the requirement of expensive, high-voltage photomultiplier detectors for ranges of the order of a mile.

LED's used with optics allow narrow-beam transmission, but optics alignment is very difficult.

Finally, common light sources have much greater spatial beam spread than LED's. However, ordinary 50,000-hour lamps can be replaced by LED's for perhaps 25 years before the LED output decreases to half power.

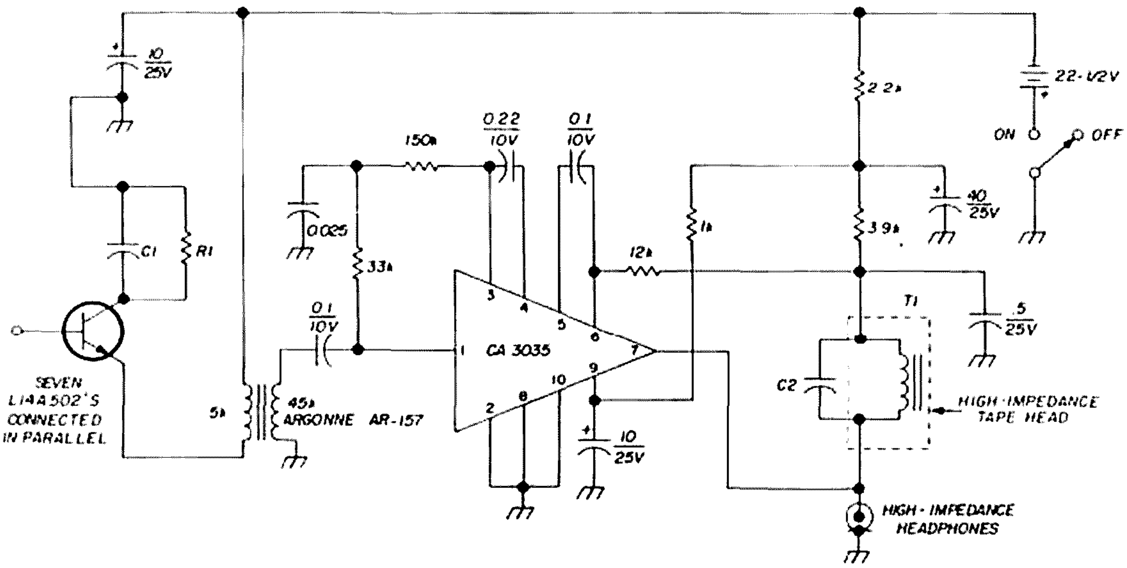



**fig. 9. Improved broad-area detector using L14A502 photo transistors and an RCA CA3035 IC audio amplifier. With this unit and an SSL-4 light source, I achieved a range of 250 feet. C1 is seven .01 capacitors in parallel; R1 is seven 680-ohm resistors in parallel. T1 is a surplus magnetic tape head. Capacitor C2 is from .01 to .033, depending upon desired audio response.**

**Another view of the broad-area detector.**

## references

1. "Solid-State Lamp Manual 3-8270," Specialty Lamps Market Development & Product Planning, Miniature Lamp Dept., General Electric Company, Nela Park, Cleveland, Ohio 44112. (Price: $2.00 for 1968 edition)
2. E. K. Howell, "Effective Irradiance," Technical Information Series R66-EGP-9, General Electric Semiconductor Products Dept., Electronics Park, Syracuse, New York.
3. Allan Lytel, "ABC's of Lasers and Masers, No. IAL-2," Howard W. Sams/Bobbs-Merrill Publishing Company, Indianapolis, Indiana, 1965. Price: $1.95.
4. "L14A502 Data Sheet," General Electric Company, Nela Park, Cleveland, Ohio 44112.
5. "Solid-State Lamp Bulletin 3-8268," Miniature Lamp Dept., General Electric Company, Nela Park, Cleveland, Ohio 44112.
6. "Application Note 3439," RCA Electronic Components & Devices, Harrison, New Jersey, issued June, 1967.

ham radio

# modulation standards

## for vhf fm

A discussion of modulation circuits and techniques to improve the performance of fm systems

Les Cobb, W6TEE, 4124 Pasadena Avenue, Sacramento, California 95821

**It is not widely realized** that there are certain variables in frequency modulation or fm that must be defined and standardized before full compatibility is obtained between transmitting and receiving equipment. In this article I will attempt to identify these variables, point out current standard practice and discuss how these standards affect transmitter and receiver circuitry.

### modulation level

In amplitude modulation systems the modulation limit is related to carrier level. This limit is called 100 percent modulation. There is no such inherent limitation for fm systems. Any modulation level, or deviation, may be transmitted as long as the receiver bandwidth will accept it.

Two standard receiver bandwidths are currently found in amateur practice. These bandwidths, as well as most of the other standards which we will discuss, stem from commercial practice—and the large amount of commercial fm equipment used by amateurs. The most common bandwidth permits a deviation of ±15 kHz; this referred to as wideband. Newer commercial equipment permits a deviation of only ±5 kHz; this referred to as narrowband. (Narrowband should not be confused with the nbfm permitted on the amateur bands below 30 MHz; nbfm is limited by regulation to ±3 kHz.)

Narrowband may be copied on a wideband receiver with only a slight loss of audio, but wideband is not copyable on a narrowband receiver because of modulation excursions out of the receiver passband. When both types of equipment are in use, modulation levels are set for the narrower receivers.

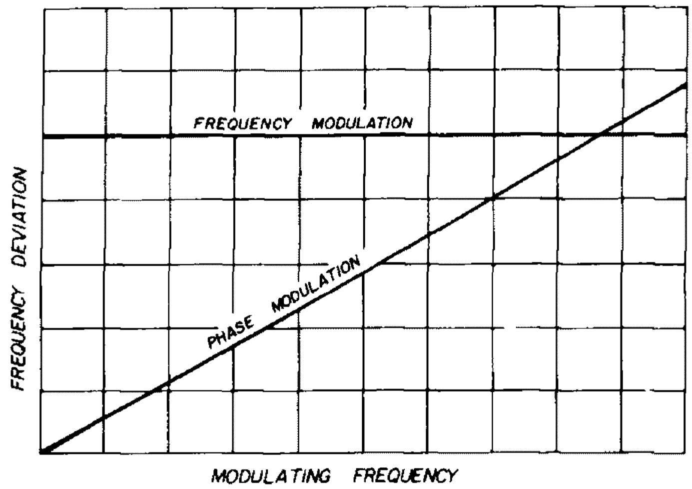



fig. 1. Modulating frequency dependence of fm and pm with constant audio input level.

## audio response characteristics

Through use, "frequency modulation" has come to refer to any angular modulation system, either true fm or pm (phase modulation). Although the difference between an fm and a pm modulator is known, it is not widely realized that the two systems result in an inherent difference in audio-response characteristics.

A constant audio level applied to a frequency modulator will result in a certain frequency deviation which does not change with the modulating frequency. However, a constant audio level applied to a phase modulator will only result in a constant peak phase shift. The frequency deviation depends on how rapidly the phase shifts. Since the phase shift becomes more rapid as the modulating frequency is increased, the frequency deviation of a phase-modulated transmitter is directly proportional to the modulating frequency as shown in **fig. 1**.

The result is that a pm signal detected in an fm discriminator will have a 6 dB per octave rising audio characteristic. This can be overcome in one of two ways. If an RC network that will cause a 6-dB-per-octave *rolloff* across the entire audio range is placed in the transmitter audio (before the phase modulator) the transmitted signal will be identical to a true fm signal **(fig. 2B)**. The alternative is to place the same RC circuit after the fm discriminator in the receiver **(fig. 3A)**. In this case

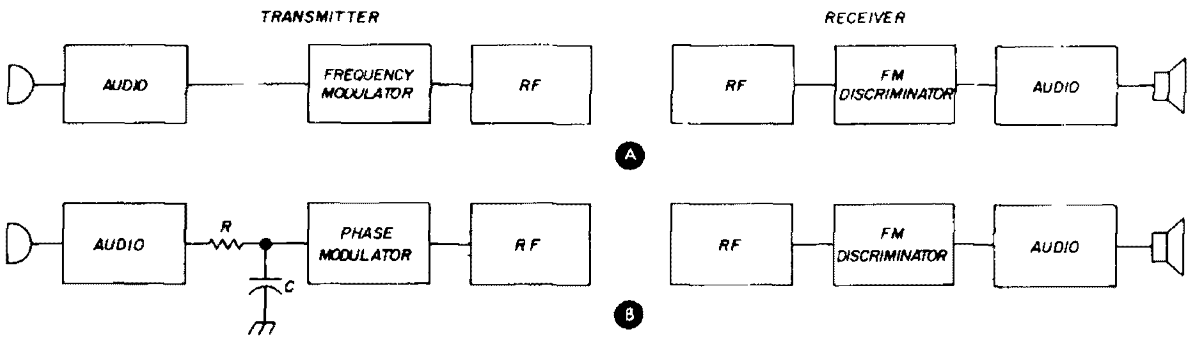



fig. 2. Frequency modulation and fm-equivalent systems.

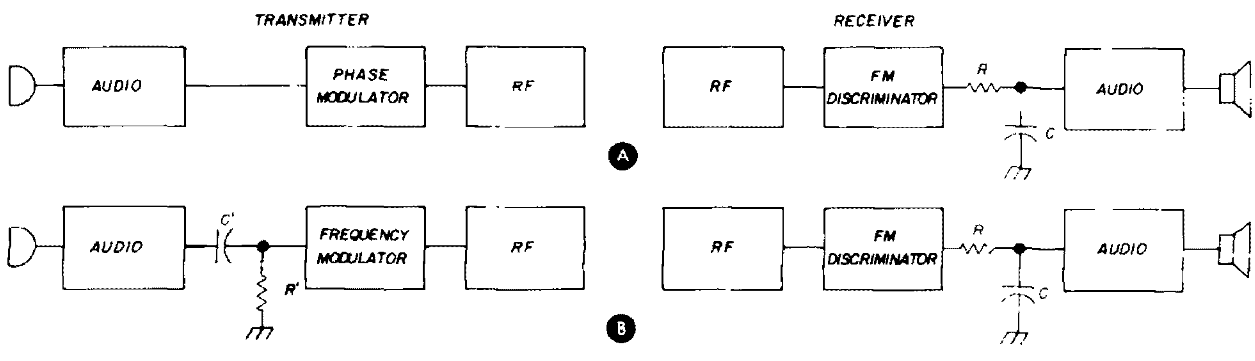



fig. 3. Phase modulation and pm-equivalent system.

the system audio will still have a net flat response but the transmitted signal will be pm.

It is pm which is standard for commercial[1,2] and amateur use. For this reason, when a frequency modulator is used an RC network with a 6-dB-per-octave *rising* characteristic is placed in the transmit audio circuit prior to the modulator (**fig. 3B**). If steps are not taken to assure standardized audio response different equipment combinations can result in either high- or low-pitched received audio with accompanying loss in intelligibility.

The RC rolloff network used in the above examples should have a time constant of RC=530 microseconds for a low-frequency limit of 300 Hz. The rising

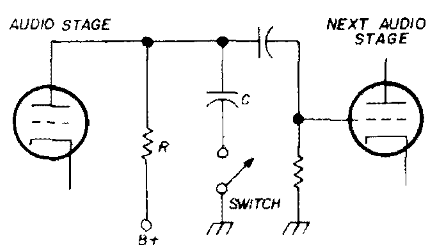



**fig. 4. Modifying an a-m receiver to slope detect pm. Capacitor C and switch are added. Shunt circuit impedances are assumed to be high relative to R and are ignored in computing RC.**

response RC network for use with a frequency modulator should have a time constant of R'C'=53 microseconds for a high-frequency limit of 3 kHz (R in ohms, C in farads). The closest standard component values may be used.

An improvement in reception may be gained when slope detecting pm on an a-m receiver if audio rolloff is added as with the fm discriminator. Not only will the unnatural high pitched quality be *eliminated, some noise reduction will* result. A shunt capacitor may be selected for the proper time constant (530 microseconds may be used) in conjunction with an existing plate load resistor (see **fig. 4**).

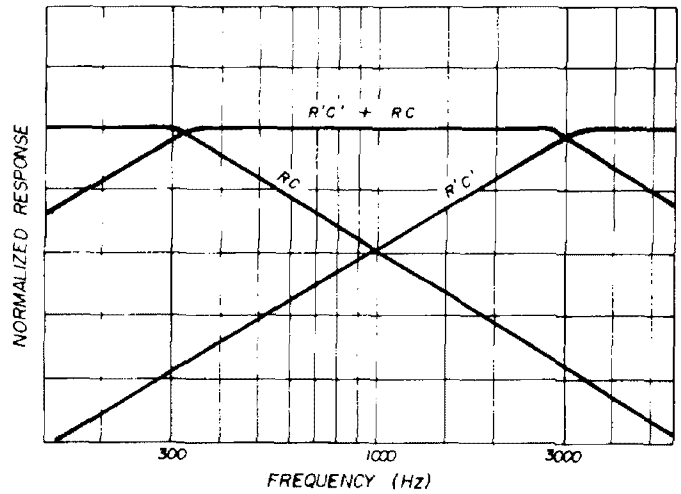


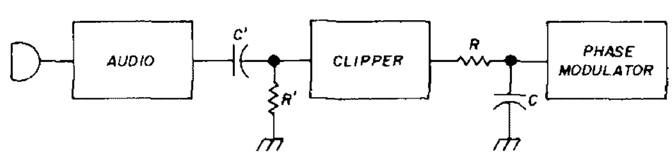

**fig. 6. Normalized response of the circuit of fig. 5.**

Provision should be made to switch the capacitor out for a-m reception. This arrangement is recommended for monitoring purposes only because of the inferior reception provided by slope detection. Also, tunable receivers are discouraged for fm communications because they encourage poor operating practices.

## speech clipping

Speech clipping is a useful method of maintaining high average deviation levels without going beyond the receiver bandpass. It has previously been established that the system in use is phase modulation; since pm exhibits a different deviation level for each modulating frequency it's obvious that fixed amplitude clipping *by itself will not work unless it is made*

**fig. 5. Speech clipping for constant maximum frequency deviation with phase modulation. R'C' = 53 microseconds and RC = 530 microseconds for 3-dB points at 300 and 3000 Hz.**



frequency dependent. This is normally done as shown in fig. 5 by preceding the clipper with a network with a 6-dB-per-octave *rising* characteristic. This enables

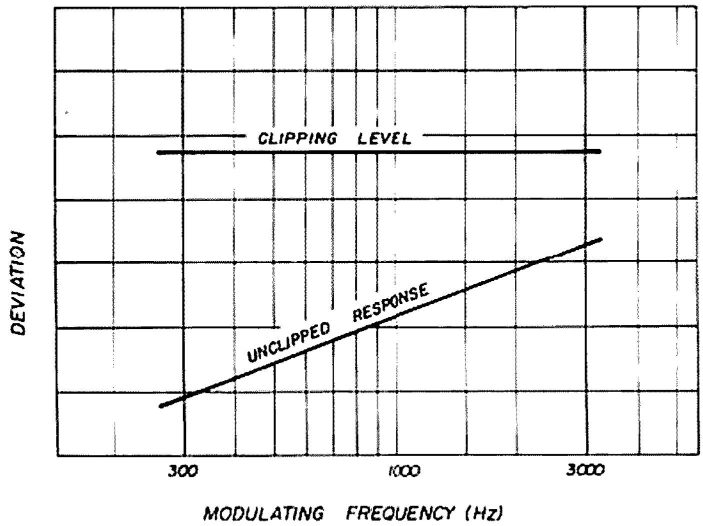



**fig. 7. Deviation characteristics of the circuit of fig. 5.**

the clipper to take a bigger bite of the higher frequencies. The clipper is followed by a 6-db-per-octave *rolloff* network that restores the unclipped audio to a flat response as shown in fig. 6. The net result is a pm signal clipped to a constant maximum frequency deviation.

When the audio clipper is used with a frequency modulator rather than a phase modulator, network RC is left out but R'C' is left in. The resulting signal is the same as pm limited to a constant maximum frequency deviation.

It should be noted that excessive clipping with this method will cause a noticeable loss of high audio frequencies. However, at normal clipping levels the spectral distribution of speech is such that little high-frequency clipping takes place, and the highs appear normal. This loss effect has been noted on many improperly adjusted repeaters around the country where the receiver is overdriving the clipper. Not only is excessive distortion created by too much clipping, but further degradation of intelligibility is caused by the muffled highs.

## summary

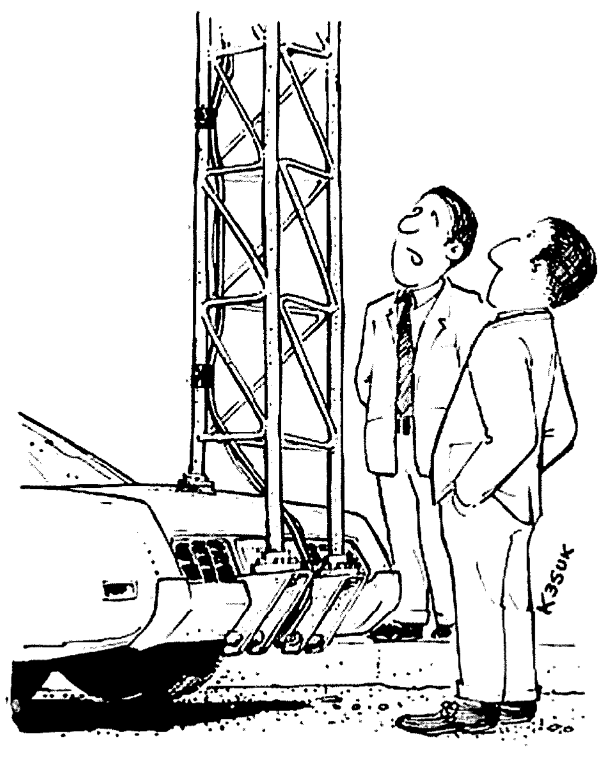
Despite the fact that fm is the general term applied to angular-modulated vhf and uhf work, the truth is that pm is the system in use from the point of view of system audio response. Audio compensation must be used with fm modulators and detectors to maintain correct audio recovery for maximum intelligibility.

Modulation levels are restricted only by receiver bandwidths (except on those lower frequencies where the FCC specifies maximum bandwidths). Speech clipping is almost universally used but special audio frequency processing is necessary in the transmitter to limit a pm signal to a constant maximum frequency deviation. Standard modulation levels are wideband (15-kHz deviation) and narrowband (5-kHz deviation).

## reference

1. EIA Standard RS-152A, "Minimum Standard for Land-Mobile Communications FM or PM Transmitters 25-470 Mc.," Electronic Industries Association, 1959, Section 6.
2. EIA Standard RS-204, "Minimum Standards for Land-Mobile Communications FM or PM Receivers," Electronic Industries Association, 1958, Section 11.

ham radio




"I'll betcha a steak dinner that he's not married."

# solid-state conversion of the gdo

Circuits for modernizing your grid-dip oscillator to obtain greater flexibility and sensitivity

Peter A. Lovelock, W6AJZ, 235 Montana Avenue, Santa Monica, California 90403

**The grid-dip oscillator** is one of the most useful items of test equipment to have around the ham station. The main shortcoming of most tube-type gdo's is their requirement for ac power. This is no problem at the workbench, but it's a definite limitation for portable or mobile work. Anyone who has used a gdo to tune an antenna knows what a chore it can be to run an ac power extension line up a tower - not to mention the safety hazard.

Today's catalogues offer a selection of solid-state "dippers" in an attractive price range. They have the advantage of being usable anywhere. If you already have an older gdo, you may have considered trading it in for one of the comtemporary models, or maybe even building a solid-state unit from scratch.

A simpler and much cheaper solution is to convert your tube gdo to a solid-state circuit. If you're reluctant about tearing into a commercially built unit or kit - don't be. The conversion task is simple, painless, and can be done in an evening. The result will give you the performance and flexibility of the latest models at a fraction of the cost.

## the tuned circuit

Before you reach for the soldering iron, inspect your tube-type gdo's schematic. The tuned circuit will influence your decision on the solid-state circuit to use. You'll want to keep the tuned circuit intact as well as the dial calibration. Thus, you won't have to change your plug-in coils.

The gdo is nothing more than a simple oscillator. In tube types, the rectified grid current is measured on a meter to indicate a "dip" when power is absorbed

from a nearby resonant circuit. Solid-state devices don't have grids, or course, so an indication on a solid-state gdo's meter is obtained from the oscillator's rectified output. The basic operating principle is the same in both circuits.

Common tuned tank circuits used in commercially built gdo's are shown in **fig. 1**. Your schematic will show if your unit has a split-capacitor, parallel-grounded, or parallel-ungrounded tank. This will determine the type of solid-state circuit you can use.

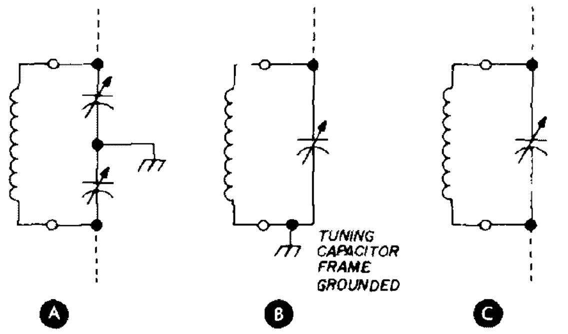



**fig. 1. Typical tuned circuits used in gdo's. Split-capacitor tank is shown in A; parallel grounded and parallel ungrounded versions in B and C.**

For the solid-state device, you have a choice of a bipolar transistor, fet, unijunction transistor, or tunnel diode. All give good performance with minor variations. For simplicity, only the first two are considered. However, if you have a favorite unijunction-diode circuit you might try it. Your final decision will probably be based on what's on hand.

## npn or pnp circuit

An npn transistor circuit I used in converting a Heath model GD-1B, which has a split-stator tank, is shown in **fig. 2.** This circuit worked well with many transistors, including the 2N2926 and 2N706, up to 200 MHz.

A pnp transistor may be used in the same circuit if you reverse the battery polarity. In both cases oscillator output was more stable than in the original tube circuit. Less frequent adjustment of the sensitivity control was required during measurements.

## common-base circuit

If your tube gdo has an ungrounded parallel tank, the common-base circuit shown on page 442 of the *RCA Transistor Manual,* Series SC-12 (reproduced in **fig. 3**), is suitable.

## fet oscillator

The circuit I finally used to convert my Heath GD-1B is shown in **fig. 4**. Advantages over the circuit in **fig. 2** are fewer components and greater sensitivity in obtaining a dip. This circuit requires a higher voltage supply, however. I used two 9-volt transistor batteries in series to obtain full-scale meter deflection over the instrument's range.

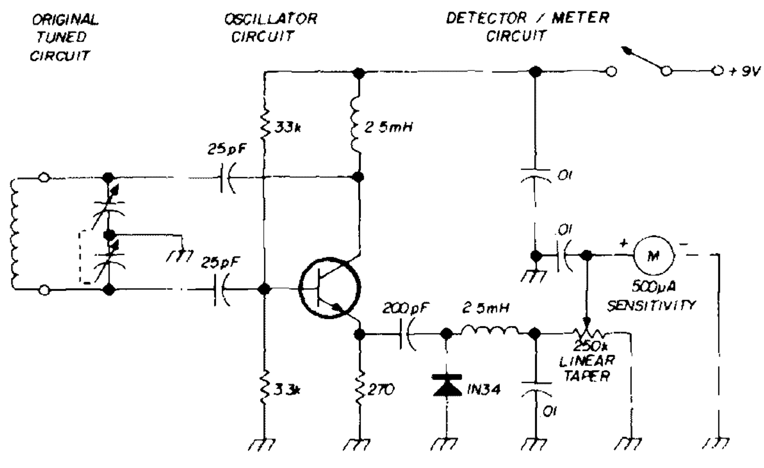



**fig. 2. Solid-state gdo with split-stator tank. A pnp transistor could also be used by reversing battery polarity.**

Since it is impractical to illustrate all

the applicable circuits for gdo conversion, I've included a list of articles in the references that should contain circuits you can use.

## construction

After you've selected a suitable circuit, you're ready to start construction. Remove all the original oscillator and power-supply components (if any) and their wiring. Don't remove the tuning *capacitor, coil socket, meter, or sensitivity* control. Take care not to disturb the wiring between the tuning capacitor and coil socket.

The logical spot for the transistor is that vacated by the tube. You can mount a transistor socket on an adapter plate placed over the tube-socket hole. If you don't like transistor sockets, cut and drill a small piece of perforated board and mount it over the tube-socket hole. Flea clips inserted in the board will allow permanent soldering of the transistor—but don't do this until all other components are mounted.

After assembling and wiring the components, temporarily attach the transistor leads to the flea clips without soldering. This allows preliminary checkout.

The photograph shows how the transistor was mounted in the Heathkit GD-1B. The socket mounting tabs were soldered directly to the copper-plated bracket that originally held the tube. Component leads must be kept short, particularly thoses connected directly to the transistor and the tuned circuit.

Small-value capacitors should be high-grade silver mica. Bypass capacitors should be ceramic, *not* paper, to avoid stray resonances in the oscillator. All resistors are composition type, ¼ or ½ watt.

The battery may be mounted in the space previously occupied by the power supply, using an appropriate bracket for the type of battery suited to your voltage and space requirements. Be sure to wire

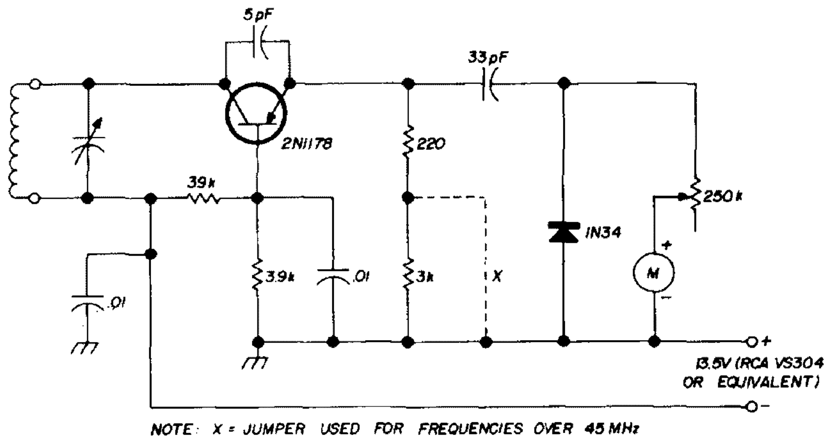



**fig. 3. Common-base gdo circuit reproduced from the RCA Transistor Manual.**

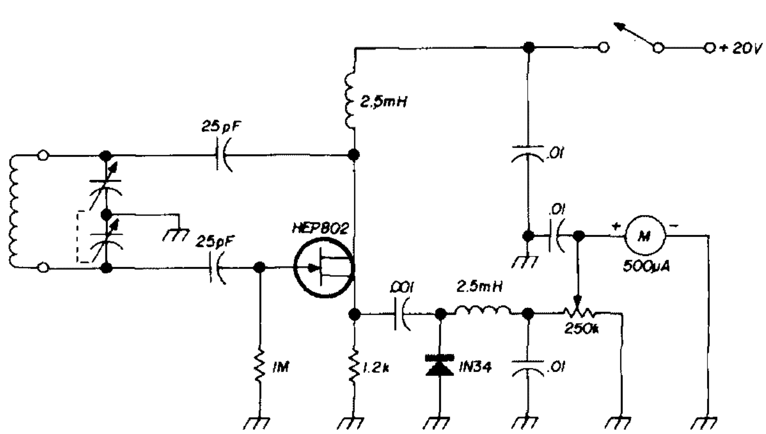



**fig. 4. Grid-dip oscillator using an fet. This circuit provides greater sensitivity with less coupling because of fet's high input impedance.**

the battery connector with the *correct* polarity for npn or pnp transistors.

In the circuits shown in fig. 2 and 4 the sensitivity control is a 250k, linear-taper potentiometer. If your gdo uses a lower value, I suggest replacing it with a 250k potentiometer and an spst switch to control battery power.

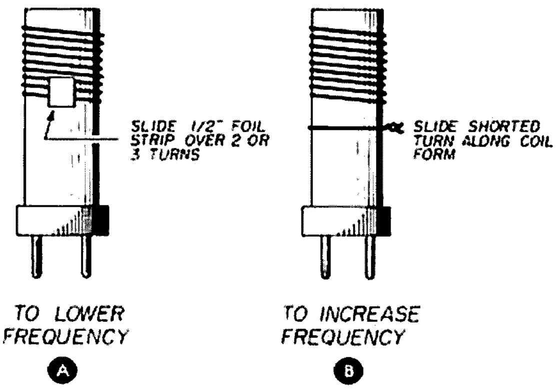



**fig. 5. Methods for adjusting gdo coils for calibration correction.**

## checkout

After wiring and carefully checking the circuit, install the battery and transistor. Plug in a coil, apply power, and turn up the sensitivity control. If you don't get a meter reading, the circuit isn't oscillating or you forgot to use a heat sink when soldering the diode rectifier.

Assuming you obtain a reading, increase the control for full-scale meter indication and tune the capacitor from minimum to maximum to check for full-scale readings over the entire range. Repeat this for each coil. If any false dips are noted without the coil coupled to another circuit, you have a "built-in" resonance. Most likely this will occur on the higher-frequency coils (40 to 200 MHz) if lead lengths are too long or if nonresonant bypass capacitors were used.

## calibration

Finally, check the dial calibration by beating the oscillator against a good communications receiver. Calibration may be a bit off if stray capacitances of the new circuit vary from the original. While most dippers are only approximately calibrate, you'll want to maintain reasonably accurate calibration. Loosening the dial-locking screw and readjusting its position relative to the tuning capacitor will take care of most cases. However, if the calibration error exceeds this method of correction, or if the error occurs only on certain coils, the following tips will help.

Sliding a one-half inch strip of aluminum foil over two or three turns of the coil will lower its frequency. Conversely, a single shorted turn of wire placed around the form will increase the coil's frequency as you slide it toward the coil. Fig. 5 illustrates these methods. After calibration has been adjusted, the shorted turn or foil strip may be permanently cemented in place.

## references

1. L. G. McCoy, W1ICP, "A Field-Effect Transistor Dipper", *QST*, February, 1968.

2. Calvin Sondergoth, W9ZTK, "Transistor Oscillators," *73*, March, 1969.

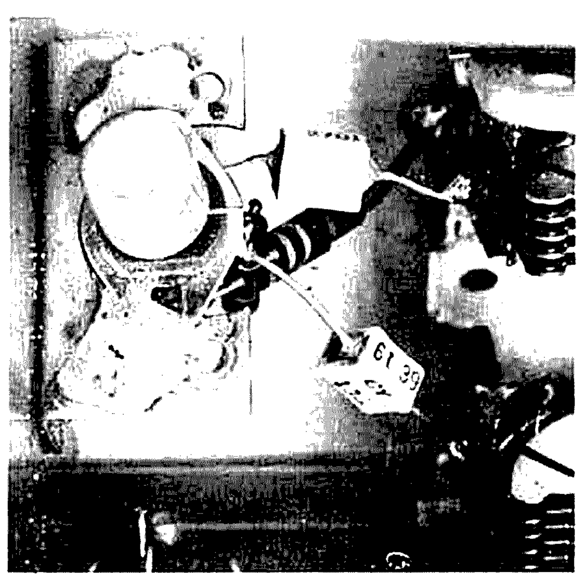
3. J. R. Fisk, W1DTY, "Designing Transistor Oscillators," *73*, August, 1969.

4. "Transistor Oscillators," *The Radio Amateur's Handbook*, ARRL Staff, 1968, Chapter 4, p. 87.

5. Rufus P. Turner, "How To Use Grid-Dip Oscillators," John F. Rider, Inc., New York, N.Y., 1960.

ham radio

**Method of mounting transistor in the GD-1B gdo.**



# integrated audio filter-frequency translator for cw reception

A sharp audio filter combined with a keyed audio oscillator— using economical ICs

John J. Schultz, W2EEY, 40 Rossie Street, Mystic, Connecticut 06355

Many amateurs like to use an audio filter to improve selectivity when receiving cw signals. Such a filter requires no internal modifications to a receiver, and the filter can be switched in or out of the circuit as desired. However, when using an audio filter you must accept the fact that the cw signals will have the same tone. This can become tiresome over a long period, because the audio tone has no harmonic content to provide a more pleasing musical quality.

An audio filter should have sharp response to be effective. Really sharp audio filters within the useful audio-frequency range are expensive, except for some surplus types such as the FL-8. Also, since filter response must be in the 800-1400 Hz range, many good bargains in very sharp filters, such as the teletype units, must be disregarded.

## the audio keyer

An audio keyer has been the classic solution to the problems of (a) varying the tone frequency of the audio selective device and (b) allowing the use of filters of almost any audio frequency. The keyer operates as follows.

Receiver audio is passed through an audio filter whose output activates a keyer circuit. The keyer circuit switches the output of an audio oscillator at the same speed as the received signals. The audio oscillator tone can be varied without affecting the received signal.

Several audio-activated keyers using vacuum tubes have been described. Unfortunately, because of the components then available, these keyers were quite bulky (almost the same size as some complete receivers) and expensive. Although their advantages were recognized, it's doubtful if many amateurs attempted

to build them.

Simple integrated circuits now available allow an audio-activated keyer to be constructed very compactly and inexpensively. In fact, an audio keyer can be made to fit inside the spare space in many audio filter enclosures. The IC keyer to be described doesn't have all the refinements of the vacuum-tube unit, but it satisfies most operational needs.

stage.) The squarer stage output is fed into an enable gate, which controls the audio oscillator signal to the audio amplifier. When a square wave is present at the squarer output, the audio oscillator signal is gated to the audio amplifier. Thus, the audio oscillator follows input-signal keying to the audio filter.

Noise can also activate the stages, so a level control for the squarer is included.

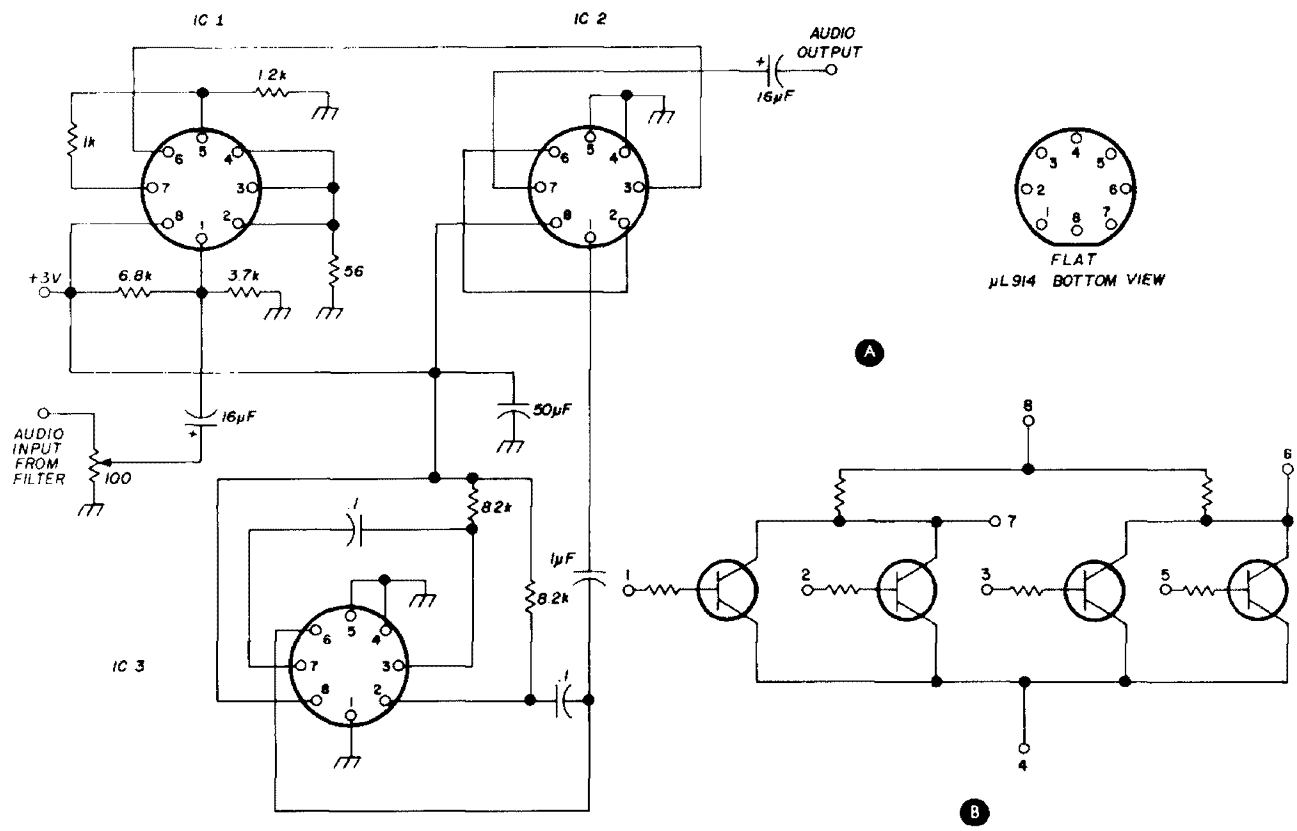



**fig. 1. Schematic of the frequency translator-keyer, A, using Fairchild μL914 IC's. The μL914 internal circuit is shown in B.**

## circuit functions

Fig. 2 is a block diagram of an IC frequency translator-keyer. The circuit differs slightly from the vacuum tube concept. Whereas the tube unit uses relay driver stages and relays for control, the IC keyer employs a wave shaper and a gating circuit.

The block functions are as follows: the squarer stage accepts the output of the audio filter, which is fed from the receiver audio output. The squarer converts the filter's sine-wave output into a square wave of the same frequency. (The squarer may be considered as a hard limiter

The time constants for coupling to the stage can be chosen to further increase noise immunity. A bypass switch is included to disable the keyer when scanning a band. It's easier to find signals without audio selectivity; also false triggering from noise and interference is avoided.

Audio filters have rather high attenuation, so the audio signal should be taken from the receiver speaker terminals rather than from the headphone jack. A transformer may or may not be required to match the audio filter input, depending on its impedance. The filter can peak at

almost any frequency as far as the keyer circuit is concerned. However, because of the restricted i-f, bfo and af response of most receivers, it's advisable to choose an audio filter in the 300- to 400-Hz range. This range is broad enough to include most audio and teletype filters described in amateur publications.

## circuit description

**Fig. 1** shows the keyer circuit using three $\mu$L914 IC's. Design has been kept as simple as possible. The input unit, IC1, is the sine-to-square wave converter. The input level potentiometer is shown as 100 ohms, but its value should be close to the audio filter output impedance. The coupling capacitor from the potentiometer arm to terminal 1 of IC1 was chosen to provide optimum performance and to avoid false triggering on noise.

IC2 is the enable gate. If the internal connections of IC2 are followed **(fig. 2A)**, it will be seen that when terminal 3 is at a positive level, the transistor associated with this terminal switches its collector to near ground potential. This places the emitter-collector resistance of the transistor associated with terminal 2 at a high level, which allows signal flow from terminal 1 to 7. When terminal 3 is not positive, the base-emitter forward bias on the transistor associated with terminal 2 rises to about the same value as the supply voltage, and terminal 7 is shorted to ground.

IC3 is a simple multivibrator audio oscillator similar to those used as monitors in electronic keyers. The components shown will provide a fundamental signal, high in harmonic content, of about 1 kHz. The resistors can be replaced by a dual 20 kilohm or 50 kilohm potentiometer if a variable tone is desired. This is especially recommended for those who like to change the receiver bfo pitch when receiving signals without an audio filter.

IC2's output is at a very low level. In general, it can be only directly coupled and used with sensitive headphones. No additional audio amplifier circuits are shown, as individual circumstances will dictate what is necessary. Any phone-type transistor amplifier can be used to boost the output of IC2.

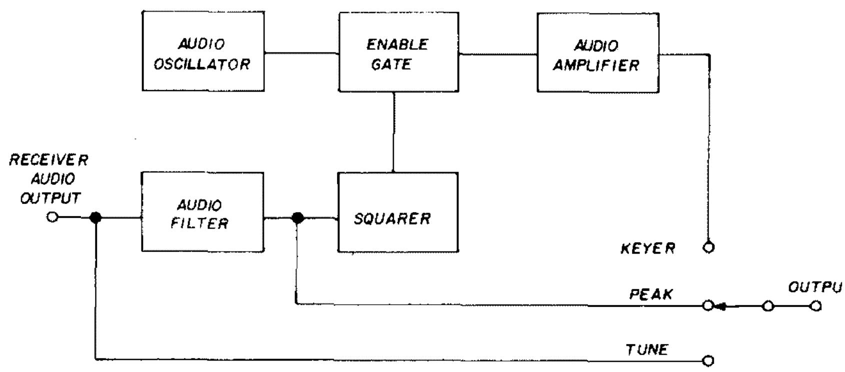



**fig. 2. Block diagram of the IC frequency translator-keyer.**

## construction

There's nothing critical about the construction or wiring. Leads should be kept reasonably short, and the wires to terminals 3 and 7 of IC2 should be separated from each other and from the connection to terminal 1.

The circuit of **fig. 1** is mounted on a vector board. Sockets *aren't* used. The $\mu$L914 IC's are soldered in place.

Supply voltage of 3-3.5 volts can be obtained from either two size-C cells in series or from a well-filtered (minimum 1,000 $\mu$F output capacitance) source within the receiver.

## operation

Tuning is done with the audio filter and keyer out of the circuit. When a

desired station is found, the mode switch is set for the audio filter output (peak position), and the receiver bfo is varied to peak the signal within the audio filter passband. Then the keyer output is chosen, and the audio oscillator frequency is varied to obtain the desired tone.

With some practice, varying receiver audio output level and keyer input level will reduce the effects of false triggering. However, if good noise immunity can't be obtained, even when using a noise limiter, the input stage (IC1) can be converted to a Schmitt trigger as shown in **fig. 3.**

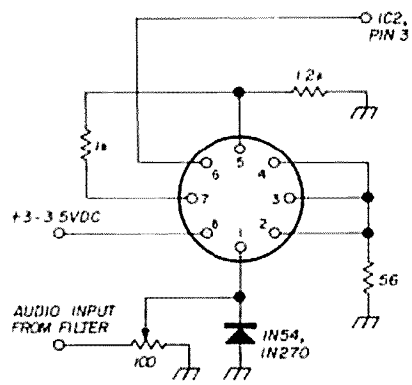



**fig. 3. Schmitt trigger circuit modification for the IC1 stage of fig. 1A involves only a simple bias change at terminal 1. The diode may also improve operation.**

Unlike the simple squarer circuit, which reacts to low-level signals, the Schmitt trigger will produce an output signal only when the input level exceeds a certain value. Thus, selection is provided against low-level noise (low in the sense of being some value less than the triggering level). The output signal will drop to zero only when the input signal falls below the triggering level.

The Schmitt trigger circuit should be used only when the noise can't be handled by the simple squarer circuit. Also, the Schmitt trigger should be used with cw agc, since large variations in audio output level between cw characters can cause the trigger to lock on or off faster than the keyer input-level control can be adjusted.

ham radio

# rf power detecting devices

What to expect from bolometers, barretters, and thermistors in measurement applications

Bruce Clark, K6JYO, 1019 El Dorado Drive, Fullerton, California 92632

**Devices for detecting electromagnetic** energy have been around since the early 1900's. These include the early electrolytic detector, carbon coherer, Fleming valve, and galena-catwhisker combination. Detectors similar to the latter are still in use. Modern materials and packaging, however, have improved their performance and reliability.

In this article I'll discuss various detectors and compare their advantages and disadvantages as applied to rf power measurement at amateur frequencies.

## basic detectors

Two types of detector are in common use. The first depends on unidirectional resistance between elements. This type includes all semiconductor diodes and their vacuum-tube counterparts. The second type changes resistance when it absorbs rf energy. Examples of this detector are bolometers, barretters, and thermistors.

Both detector types have many applications in detection and demodulation of rf energy.

## definitions

Many take for granted that detection and demodulation are one and the same thing. This is not the case. Radiofrequency energy doesn't have to carry modulated intelligence to be detected. An example is the simple diode wavemeter, which has an indicator that reveals the presence of an unmodulated rf carrier.

## demodulation

Demodulation is a byproduct of the detection process. Demodulation translates modulated intelligence, riding on the detected carrier, into a form that can be displayed aurally or visually. An example of demodulation is when an rf carrier, modulated by a 1-kHz tone, appears as a 1-kHz voltage at the detector's output.

Common diode detectors are used to convert rf energy to dc voltages and modulated rf energy to ac voltages that can be measured with orginary test equipment. Thus, rf voltages, current, or power can be quantified by using rf detectors and the proper readout device.

## the diode detector

Diode detectors have characteristics that affect their accuracy in rf-measuring applications. An understanding of their drawbacks as well as their advantages is essential to using them effectively.

Diodes have a high impedance between their elements when reverse biased, **fig. 1A,** and a low impedance when biased in the forward direction, **fig. 1B.** This resistance characteristic allows the current to flow in only one direction, resulting in dc pulses flowing in the load.

The resistance characteristic is nonlinear. It obeys a square-law function,

which occurs when the diode's output is proportional to the square of the input (assuming constant input impedance for a given range of input levels).

## equivalent circuit

Another characteristic is that the resistance is directly proportional to the power dissipated. The square-law characteristic is due to the physical properties of the junction, **fig. 2**. In this circuit, C is the barrier capacitance due to charge storage in the barrier region. R represents the nonlinear barrier resistance, which is about 5k ohms at low current levels and

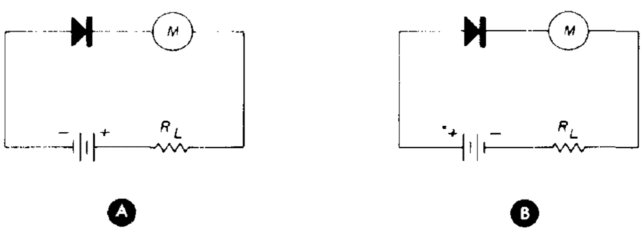



**fig. 1. Reverse- and forward-biased diode junctions. Meter response is a function of diode junction resistance, which is high when reverse biased, A, and low when forward biased, B.**

falls rapidly with increasing current. The barrier-spreading resistance, r, is due to the construction of current paths in the barrier region. Its value will be as high as 50 ohms in point-contact diodes, decreasing to tenths of an ohm in diffused-junction types.

## measurement errors

Most power and vswr meters are calibrated in terms of the square-law function. However, if too much power is applied to the diode, a readout error can result.

The diode's square-law region is typically limited to input levels between 0.25 and 1 mW. Above 1 mW, the diode's output will deviate in excess of 10 percent from the square-law response. In terms of dB error,

$$E = 10 \log (1 + d)$$

where E is the error in dB and D (100) is the percent deviation from the square-law function.

Crystal diode sensitivities of the order of 5k $\mu$V/mW are readily obtained at frequencies below 1 GHz. By proper choice of detector load resistance, the square-law range (5 percent deviation-error band) can be optimized to allow power ration (attenuation) measurements over a range of 36-38 dB with errors of the order of 0.2 dB. If the diode is overdriven, errors of 1 dB or more can occur. This is a considerable error at microwave frequencies in some applications.

## bolometers

These detectors **(fig. 3)** use fine platinum wires (known as Wollaston wires) as active elements. The active elements are small compared to a wavelength at the measurement frequency. Their changes in resistance, however, are sufficiently large for accurate measurements with small changes in input power.

Bolometers are used with a specific "bias" current, which activates a bridge circuit. Typically, a bolometer presents 200 ohms to the bridge, although this may range from 50 to 400 ohms. The bolometer generally presents a 50-ohm resistance to its input. Excitation currents are of the order of 4.5 to 4.7 mA. A simplified bridge circuit is shown in **fig. 4.**

## sensitivity

Bolometers are less sensitive than diodes. Normal sensitivities range from 4 to

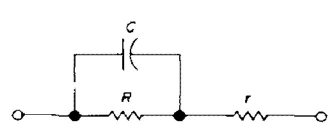



**fig. 2. Equivalent circuit of forward-biased semi-conductor diode junction. Functions of circuit constants are discussed in the text.**

10 ohms per mW of rf, depending on the temperature coefficient of the resistive elements. In terms of voltage, this works out to about 21 $\mu$V/mW.

Bolometers are preferable to diodes

when overdriving and burnout is a problem. At powers over 1 mW, bolometers deviate about 10 percent from square-law response due to convective cooling of the resistance element. However, this deviation is nearly linear with power level and is easily compensated in measurements.

This deviation from square-law response falls to less than 2 percent at 200 mW; thus bolometers can be used for relative power and attenuation measurements over a 53-dB range with less than 0.2-dB error (5 percent square-law deviation). This is approximately 15 dB more power range than is available with diodes.

The dynamic range numbers (dB) are given on the assumption that signal-to-noise ratios are at least 10, which is an acceptable value for accurate results with most instruments.

Unlike diodes, bolometers are not affected by overload up to about 32 mW, including 15 mW bias power.

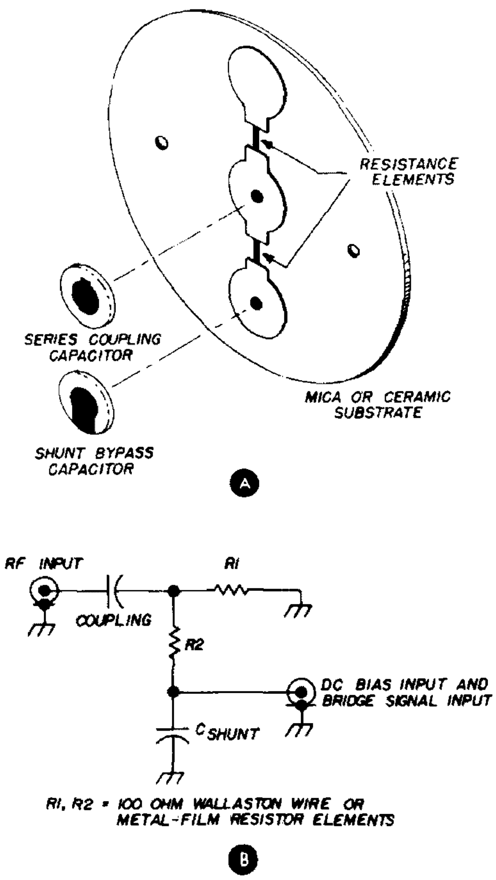



**fig. 3. Schematic, B, and physical representation, A, of bolometer. Input resistance is 50 ohms nonreactive; bias input to bridge is typically 200 ohms.**

## barretters

Barretters are a type of bolometer employing a single Wollaston wire element, **fig. 5**. Usually they're used as shunt elements, **fig. 6**, and are packaged to replace crystals in applications where wider measurement range and greater power-handling capability are desired, and where a decrease in sensitivity can be tolerated.

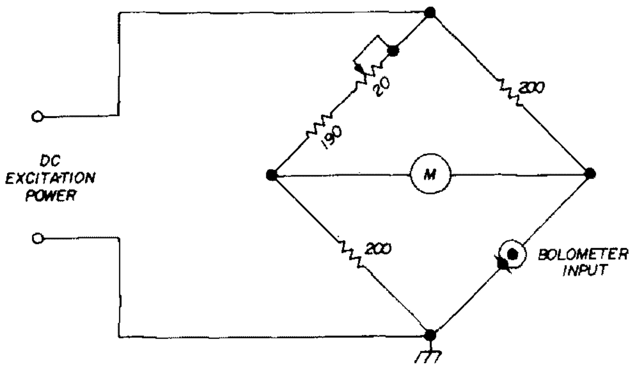



**fig. 4. Simplified bolometer or thermistor bridge circuit.**

## thermistor detector

A thermistor is another important rf detector. Thermistors are resistance elements made from one of several metallic oxides such as oxide of manganese, nickel, titanium and zinc. All exhibit a negative temperature coefficient. The slope of resistance versus temperature of thermistors is much greater than that of bolometers. Thus, thermisiors have improved sensitivities that range from 50 to 100 μV/mW of rf input.

Like bolometers, thermistors require a bias current. This is usually adjusted to provide a "zero power" resistance of 100 to 200 ohms. Thermistors also have a large overload capability and the widest dynamic range of all the rf detectors. This makes them ideally suited for accurate, steady-state power measurements. However, thermistors have disadvantages that limit their applicability. The prime limitation is *their relatively slow time constant.* Ranging between 1 and 3 seconds, it

prevents the thermistor from measuring peak pulsed or peak ssb power. Additionally, audio-frequency modulated signals can't be demodulated with a thermistor, as its resistance changes too slowly to follow an audio signal. Another disadvantage is its sensitivity to ambient temperature changes. This tends to make long-term stability a serious problem. Most thermistor rf detectors use dual elements that tend to compensate each other for drift.

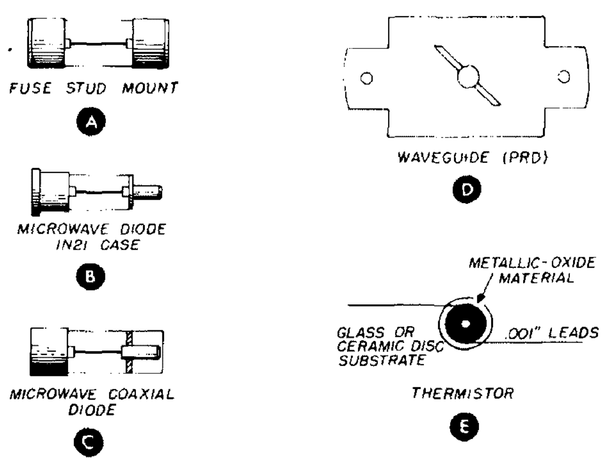



**fig. 5. Packaging of barretters and thermistors. A, B and C use Wollaston-wire elements.**

Unlike diodes and bolometers, thermistors are insensitive to overload and can be exposed to fairly high-level rf inputs for short periods without damage. The output characteristic of thermistors, although opposite in polarity, is similar in slope to the curve of bolometers and will permit a measurement range from approximately 50 µW to 5 mW.

## rf detector readouts

In many amateur applications, only a relative indication of rf power is needed. This requires only a milliammeter and a potentiometer for a sensitivity adjustment. In many other applications such as measuring transmitter efficiency, antenna gain, feedline or filter losses, or low level injection powers, it's desirable or even necessary to get a quantitative figure in watts or dB. In these cases more sophisticated circuitry must be employed. For best accuracy and repeatability, the null-balance measurement system is usually employed. Usually a variation of the Wheatstone bridge provides high accuracy especially at low power levels. Temperature stability and dynamic range are a problem with these units. Many techniques have been imployed to eliminate these problems.

## self-balanced bridge

Practical measurement bridges differ from the simple Wheatstone circuit of **fig. 4.** Usually some sort of a "self- balancing" circuit is employed.[1] Shown in **fig. 7,** the self-balancing bridge is direct reading and is much less sensitive to ambient temperature variations. With the proper bias adjustments, it can be used with all types of bolometer, barretter and thermistor detectors for input powers up to 100 mW depending on the type of detector employed. In the self-balancing circuit, the Wheatstone bridge forms a coupling network in the feedback loop of a high-gain audio amplifier. The resulting audio oscillator automatically adjusts its output voltage to maintain a balanced condition in the bridge network. When the bridge is unbalanced by an rf input, an equal amount of audio power is removed from the bridge, restoring balance. This is read out in terms of voltage by the audio voltmeter **(fig. 7)** whose scale is calibrated directly in terms of rf power.

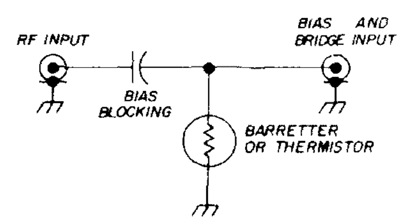



**fig. 6. Barretter or thermistor mount circuit.**

## attenuators and filters

The power range of all detectors may, of course, be greatly extended by calibrated attenuators2 and directional couplers.[3] Most ham-type swr bridges use a crude form of directional coupler to

sample relative rf power on a transmission line. Where fairly accurate results are desired, precision couplers are employed, sometimes with calibrated attenuators when even more power reduction is required. For amateur use, attenuators can be made from carbon composition resistors or from calibrated lengths of RG-58/U coax (at vhf).

Measurement errors due to improper detector use are varied. Some are of little significance in amateur applications. For amateurs, most problems result from gross overdrive of the detector. This results in destruction of the unit through burnout, or a serious departure from the idealized square-law output curve with a resultant measurement error. Errors due to this can be reduced by using attenuators and by knowing the input parameters of the detector.

Another common error resulting in amateur rf detector applications is also of importance. All rf detectors described in this article are untuned; that is, they have no inherent selectivity and will accept power over a wide frequency range. Therefore, when measuring power sources with high harmonic content, the detector will sum the power in the desired frequency as well as that in its harmonics. This results in nonexistent and colossal "efficiency." What's worse, it can cause operation on an undesired or spurious frequency.

When measuring output power from sources containing harmonics, some type of selective filter with a calibrated insertion loss should be used to reject the unwanted power before it reaches the detector.

Representative filters are described in the references. Further details on rf power detectors are contained in Henney's "Radio Engineering Handbook," 1959 edition.

I hope this discussion will provide a better understanding of this basic measurement tool and encourage its use in amateur applications.

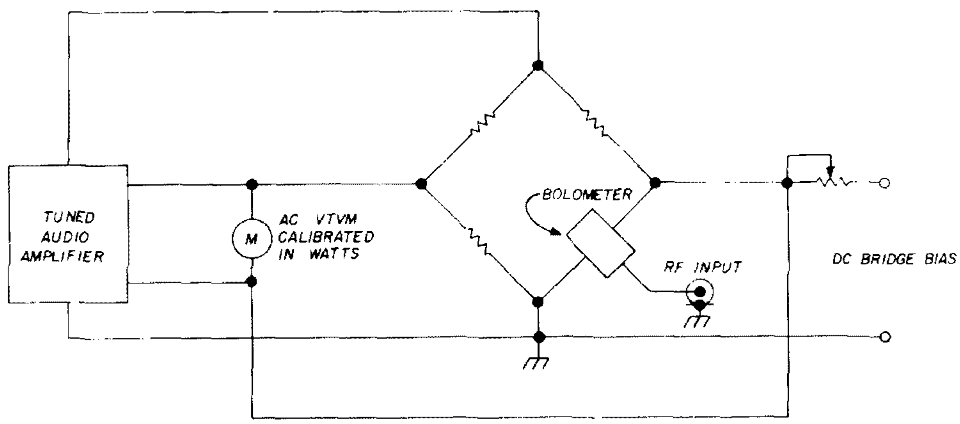



**fig. 7. Self-balancing bridge and power meter. This circuit is much less sensitive to temperature changes than the simple Wheatstone bridge.**

## references


1. *Microwave Test Equipment catalog E8,* Polytechnic Research and Development Corp., New York, New York, 1958.

2. B. Goodman, W1DX, "A Simple Step Attenuator," *QST,* August, 1969.

3. R. J. Kolb, WA6SXC, "A Coaxitube Directional Coupler", *73,* February, 1966.

4. Denton Nelson, W7UHF, "Cavity Filters," *VHF'er,* Vol. 4, No. 7, September-October, 1966, p. 5.

5. L. Baldwin, WORUG, "Simple Filters for the 1215 Band," *QST,* April, 1969.

6. R. E. Fisher, W2CQN, "Combined VHF Band Pass Filters," *QST,* December, 1968.

7. E. P. Tilton, W1HDQ, "Coaxial Filters," *VHF Handbook,* Vol. 1, ARRL.

8. F. Jones, W6AJF, "144, 220, 432 Transistor Converters," *73,* June, 1966.


**ham radio**

# design criteria for ssb phase-shift networks

How to minimize network errors for effective sideband suppression — an analysis of Norgaard and Dome circuit

A.I.H. Wade, G3NRW, 11 Daubeney Close, Harlington, Dunstable, Bedfordshire, England

**Phase-shift networks in amateur** phasing-type exciters consist of passive elements arranged to produce two equal-amplitude signals in phase quadrature. Two networks are required: one operates at audio and the other at radio frequency.

The rf network operates at a single frequency, whereas the af network must pass a band of voice frequencies. The rf network is fairly simple to adjust, but the af network is more critical as to construction and adjustment. This applies not only to the network, but to its associated circuits as well.

## theory

Two popular audio networks are analyzed: Norgaard and Dome. Design-center conditions are considered first. Then network errors are discussed in terms of their effect on carrier suppression in a practical system. A brief treatment of the rf phase-shift network is also given.

Four areas are considered:

1. The ideal case.
2. Network phase errors.
3. Network amplitude errors.
4. The general case (combined effects of phase and amplitude errors.

## ideal case

A lower-sideband signal, for example, requires a voltage of the form

$$Vo = A \cos(\omega c - \omega m)t \qquad (1)$$

where

Vo is the lower-sideband voltage

A is the signal amplitude

$\omega c$ is the carrier frequency ($\omega = 2\pi f$)

$\omega m$ is the audio-signal frequency

Two networks are required to produce this voltage, one each for the audio and carrier frequency. Each network has two outputs whose phase difference is 90°; thus four signals are produced:

a. $V1 = A \sin(\omega m\ t)$ (2)

b. $V2 = A \sin(\omega m\ t + \pi/2)$ (3)

c. $V3 = B \sin(\omega c\ t)$ (4)

d. $V4 = B \sin(\omega c\ t + \pi/2)$ (5)

where A and B are amplitudes of the audio and carrier signals respectively.

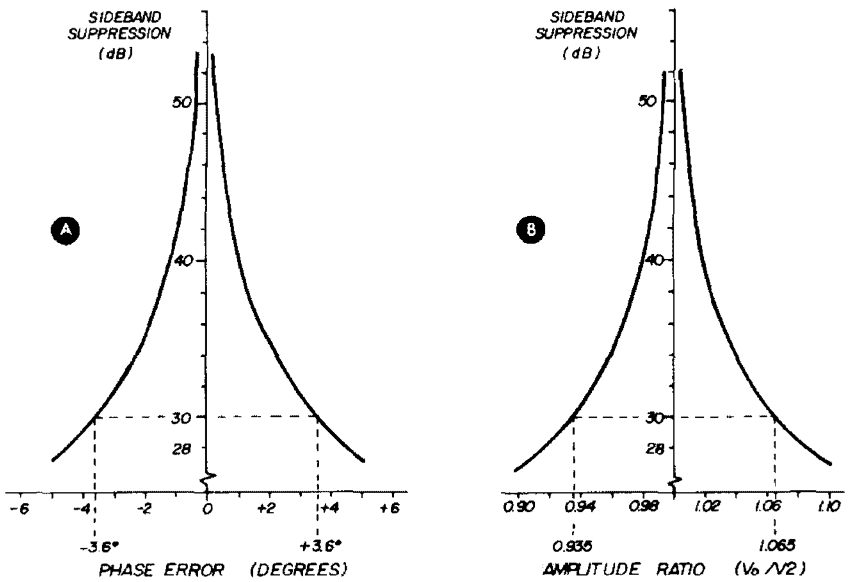



**fig. 1. Sideband suppression as a function of phase error, A, and amplitude unbalance, B, assuming ideal conditions. For 30-dB attenuation, errors must not exceed ±3.6° and 6.5 percent respectively.**

**The phasing system** of producing ssb signals hasn't been too popular. As the author points out, the audio networks are tricky to adjust and the degree of sideband suppression depends on how well the networks are designed to limit phase and amplitude errors.

Phase-shift network theory is presented here in terms of system equations. Realizing that mathematical articles don't appeal to many readers, *I feel this one is appropriate because of* the renewed interest in phase-shift techniques for direct-conversion receivers. The phasing method is also interesting for vhf ssb work. This article presents essential data for those wishing to avoid the pitfalls associated with phasing networks. With an understanding of their basic concepts, the networks can be tailored to produce quite acceptable results.

Although not included because of space limitations, an appendix accompanying the manuscript derives the equations in the text. Interested readers may obtain a copy from *ham radio* for $1.00. **editor**

If signals a and c are applied to the input of one balanced modulator, the modulator's output will be of the form

$$Va = \frac{A}{2}[\cos(\omega c - \omega m)t - \cos(\omega c + \omega m)t] \qquad (6)$$

Similarly, if signals b and c are applied to another balanced modulator, its output will be

$$Vb = \frac{A}{2}[\cos(\omega c - \omega m)t + \cos(\omega c + \omega m)t] \qquad (7)$$

At the output of each balanced modulator is a signal that contains components of both upper and lower sidebands (the carrier has, in both cases, been balanced in the modulators).

The outputs of the two balanced modulators are combined:

$$Vo = Va + Vb = A \cos (\omega c - \omega m)t \quad \textbf{(8)}$$

The upper sideband has now been cancelled leaving only the desired lower-sideband signal.

## phase errors

In the previous discussion it was assumed that each network shifted the phase through precisely 90°. This can't be done in practical networks, so phase error terms $\phi$ and $\theta$ must be introduced into **equation (3)** and **(5)** to account for this. It can be shown that the lower and upper sideband amplitudes will then become $A \cos (\phi - \theta)/2$ and $A \sin (\phi + \theta)/2$ respectively. That is, an unwanted upper sideband has been generated whose amplitude is a function of the *sum* of the phase errors of the two networks. The unwanted sideband can be eliminated if the phase errors are exactly equal in magnitude but of opposite sign. In practice, this can't be obtained over the frequency range in which the networks must operate.

Note that, if the rf network has no phase error, $\theta$ becomes zero, and the amplitudes of the two sidebands reduce to $A \cos (\phi/2)$ and $A \sin (\theta/2)$.

Defining the sideband suppression, S, as the ratio (in dB) of the lower-to-upper sideband,

$$S \text{ (in dB)} = 20 \log_{10} \cot \phi/2 \quad \textbf{(9)}$$

This is plotted in **fig. 1**A. Taking the minimum acceptable value of suppression as 30 dB, the maximum phase error that can be tolerated is of the order of ±3.6°. This assumes, however, that the rest of the system is perfect. In practice this number must be reduced.

## amplitude unbalance

In eqs. 2 and through 5, it was assumed that the af quadrature signals were of equal amplitude, A, and the two rf signals were of equal amplitude, B. Again, in practice, this condition won't be met exactly, thus requiring a further change to the four equations. To simplify matters, however, assume that the rf network can be adjusted to produce two equal-amplitude signals, and the two outputs are of magnitudes A and nA, where n is approximately unity.

An unwanted upper sideband is again generated which is of amplitude $A/2\ (n - 1)$. Its suppression is

$$S \text{ (in dB)} = 20 \log_{10} (n + 1)/(n - 1) \quad \textbf{(10)}$$

This dependence of sideband suppression on the value of n is shown in **fig. 1B**. As n approaches unity, the suppression tends to infinity; and to satisfy the limiting condition of 30 dB, n must be between 0.935 and 1.065. That is, the amplitudes of the two audio signals must not differ by more than 6.5 percent.

## general case

Up to this point we've examined separately the effects of phase errors and amplitude unbalance in the network outputs. To illustrate what happens in the general case, these effects must be combined. In this case, the sideband suppression is

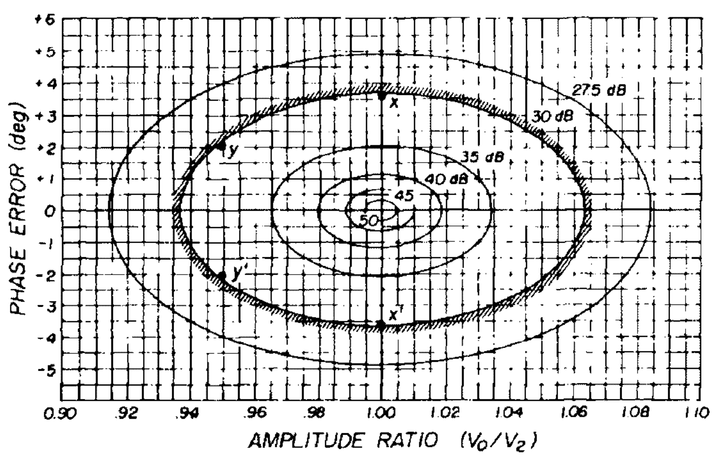



**fig. 2. Combined effects of network amplitude and phase errors. Errors must lie within shaded area for acceptable sideband suppression.**

S (in dB) =

$$10 \log_{10} \frac{1 + n^2 + 2n \cos\phi}{1 + n^2 - 2n \cos\phi} \qquad (11)$$

From this equation the curves of **fig. 2** are plotted: the suppression may be read directly once the phase error and amplitude unbalance are known. The curve corresponding to the suppression limit of 30 dB is shaded; for acceptable operation, the audio phase-shift networks errors must not lie outside this area at any frequency. For example, a phase error of ±3.6° is acceptable only with perfect amplitude balance (points X); if the amplitudes differ by, say, 5 percent, the phase error can't exceed ±2.2° (points Y).

From the foregoing we can consider the requirements of an ssb phasing exciter, a block diagram of which is shown in **fig. 3.** The amplifiers in the audio frequency chain are considered first.

## audio preamplifier

The audio preamplifier must

1. Amplify the low-level output of the microphone.

2. Overcome the insertion loss of the audio phase-shift network which follows it.

3. Tailor the frequency response so that frequencies outside the range of 300 to 3000 Hz are attenuated.

The first two requirements are easily met if usual precautions are taken to minimize distortion. For the third requirement, filtering must be used in the preamplifier because the signals presented to the phase-shift network must be restricted to the frequency range mentioned. The reason is that practical networks exhibit increasing phase errors outside this range, producing poor sideband suppression.

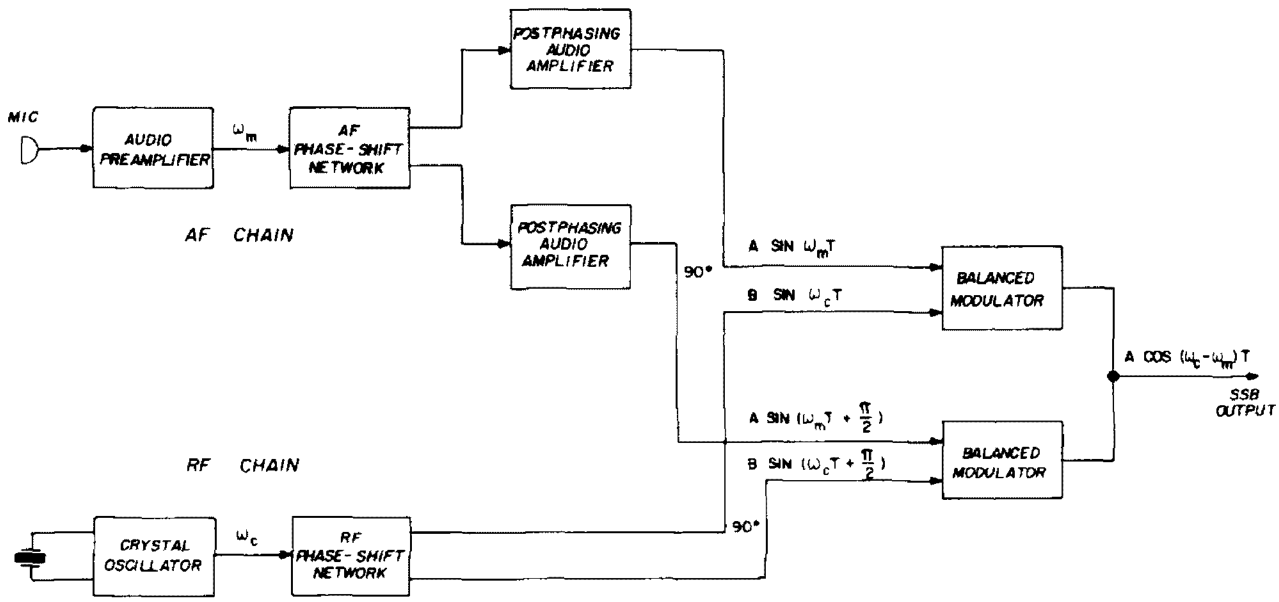



**fig. 3. Typical phasing-type ssb exciter.**

## postphasing amplifiers

Each audio phase-shift network output drives an audio amplifier. The amplifiers buffer the network outputs from the modulator inputs and offer the correct terminating impedance to the networks.

Optimum phase-shift network performance requires correct output terminating impedance, which is usually high and must be constant. The amplifiers must introduce no phase or amplitude errors in addition to those produced by

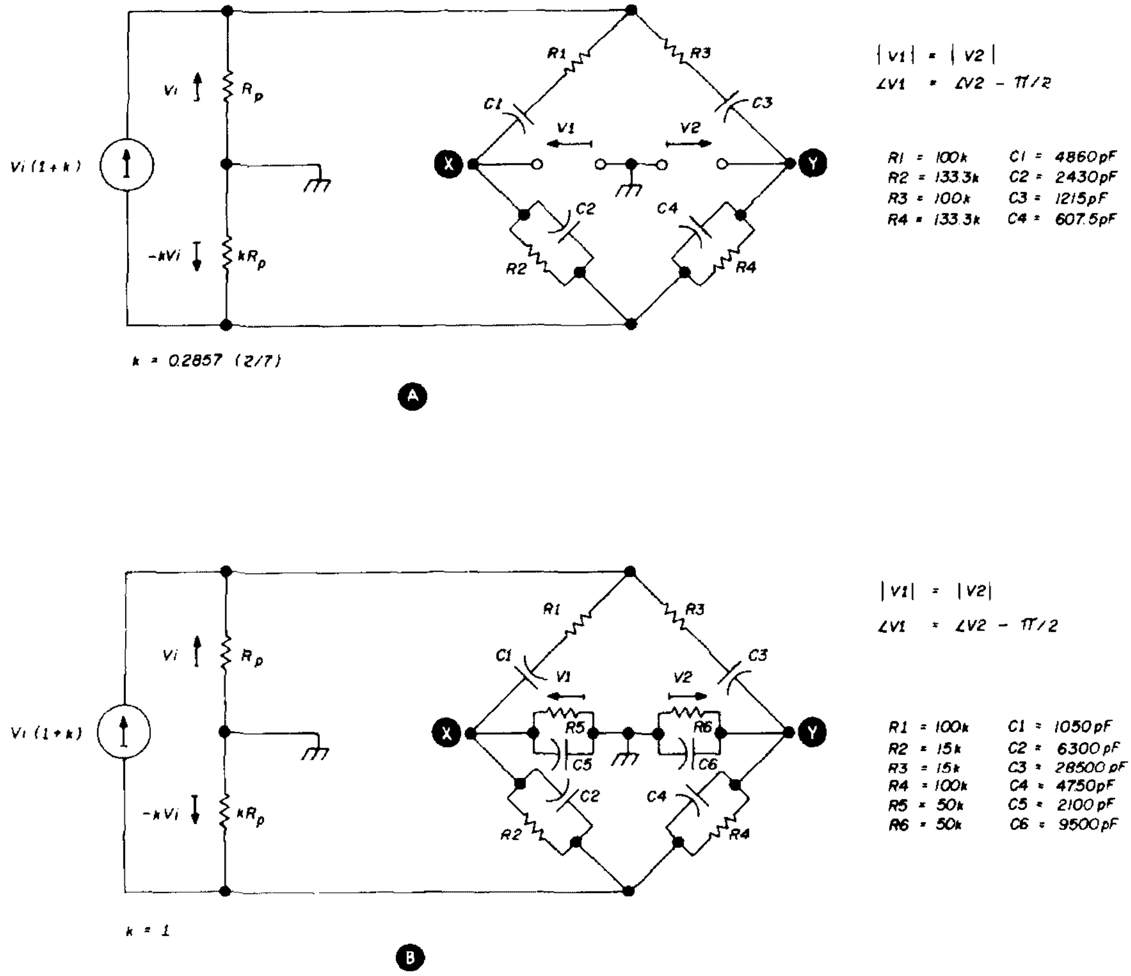



**fig. 4.** Norgaard network, A, and Dome network, B. Input drive floats in both circuits. Equal-amplitude outputs are taken from points X and Y.

the phase-shift network. These amplifiers must have a wide frequency response and minimum amplitude distortion.

## rf circuits

The rf chain of an ssb exciter functions similarly to the audio chain, but at a fixed, comparitively low frequency. This simplifies its design because

1. The oscillator frequency may be accurately maintained by crystal control.

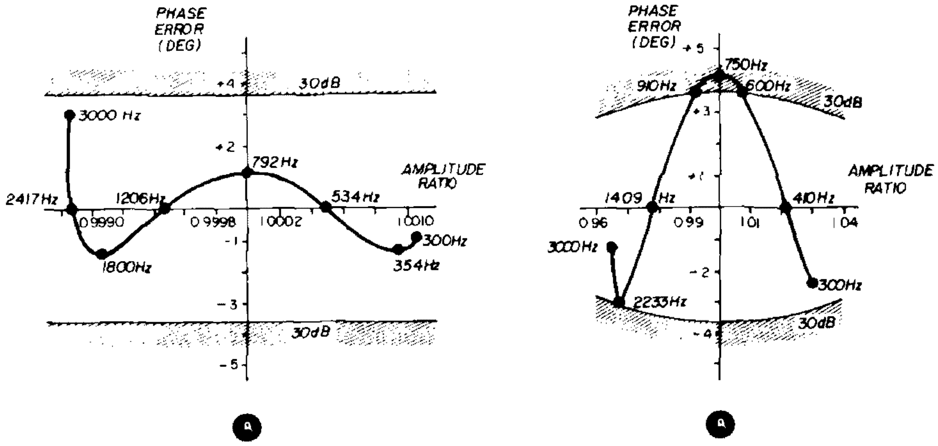



**fig. 5.** Error response of Norgaard network, A, and Dome network, B. Suppression with the Dome network falls below 30 dB between 600 and 910 Hz.

2. The rf phase-shift network operates at only one frequency, so it's simple to design, construct, and adjust.

Crystal frequency is determined by operating frequency and unwanted mixer products. The final operating frequency is obtained by mixing the ssb output of the exciter with a heterodyned frequency.

## detailed network analysis

The Norgaard and Dome audio phase-

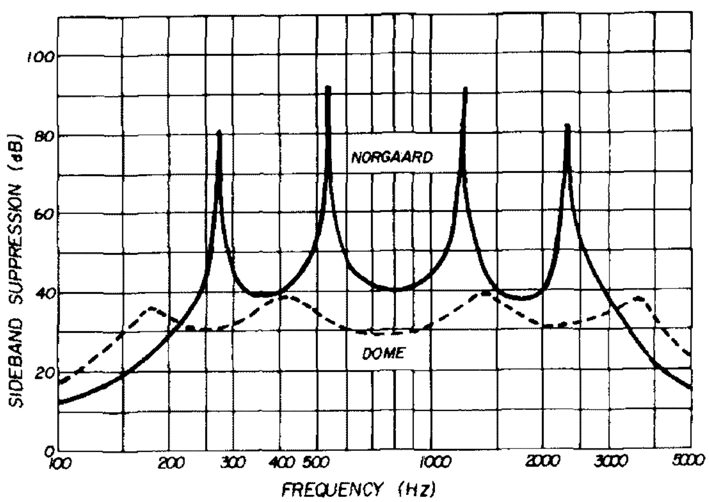



**fig. 6. Sideband suppression as a function of frequency. Note marginal performance of the Dome network around 750 Hz.**

shift networks have the following characteristics:

1. The Norgaard (**fig. 4A**) is a bridge network designed to be terminated in an infinite impedance. Its input requires two antiphase signals whose amplitudes are in the ratio 2:7.

2. The Dome is shown in **fig. 4B**. It is a full lattice network. Outputs are developed across a finite, complex impedance. Its input requires two antiphase signals of equal amplitude.

Each network operates on the differential phase principle. The phase angle with respect to the input of *each* output increases with frequency, but the phase difference between the outputs remains constant at 90°.

The amplitude and phase response of the Norgaard and Dome networks are shown in **fig. 5A** and **5B**. In both cases *ideal* component values, input drive ratios, and terminating impedances are assumed. The arrows on the curves indicate the direction of increasing frequency; several spot frequencies are also shown.

## amplitude response

In the Norgaard network, the amplitude ratio of the two outputs remains very close to unity; worst-case unbalance is of the order of only 0.1 percent. The suppression curves are quite flat in this region, so amplitude unbalance has negligible effect on the final suppression value.

In the Dome network, maximum error is approximately 3 percent. Amplitude unbalance is therefore responsible for degrading maximum obtainable suppres-

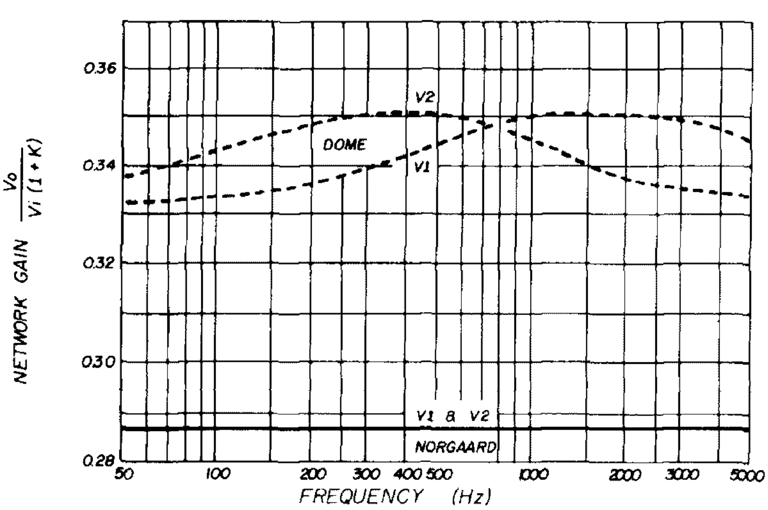



**fig. 7. Insertion loss of Norgaard and Dome phase-shift networks.**

sion by several dB.

## phase response

The maximum phase error produced by the Norgaard network is +2.9° at 3000 Hz. The phase error is primarily responsible for limiting sideband suppression.

**Fig. 6** compares the two networks over a wider frequency range. Note sideband suppression at line frequency. The Norgaard network suppresses the unwanted sideband 7.5 dB at 60 Hz, increasing to 15 dB at 120 Hz. This means that the audio preamplifier must be designed to attenuate these frequencies on the unwanted sideband.

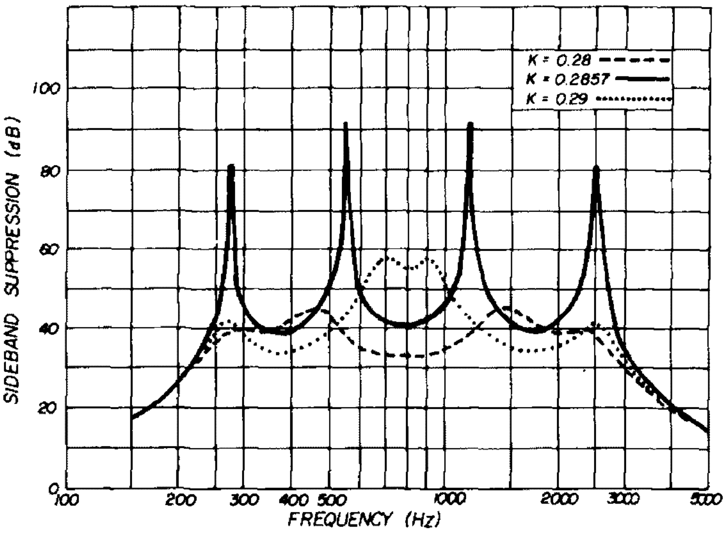
The Dome network phase error ranges from -2.9° to 4.2°. Unwanted sideband suppression decreases to 28.8 dB at 750 Hz.

## network gain

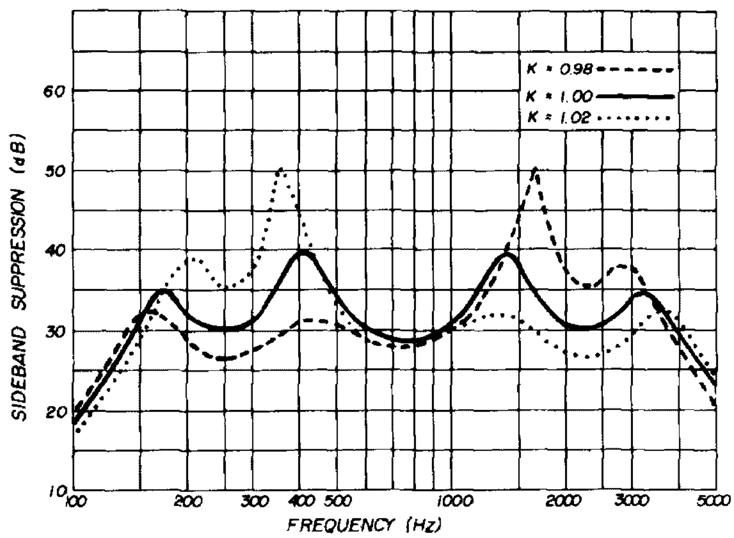
If the amplitudes of the two antiphase input signals are Vi and kVi, where k is the ratio of the appropriate network (i.e., 2/7 or 1), and Vo is the output signal amplitude, then network gain is

$$G = \frac{Vo}{Vi\,(1+k)} \qquad (12)$$

This plotted versus frequency for each network **(fig. 7)**. Note that little variation in insertion loss is evident for either network. Expressing the gain at 1 kHz (by taking Vo as the mean of the nearly equal otuputs), the Norgaard and Dome networks exhibit -13.1 and -15.2 dB respectively.




A




B

**fig. 8. Effect of *drive-ratio errors* (i.e., incorrect values of k). Norgaard network response is shown in A; Dome network in B.**

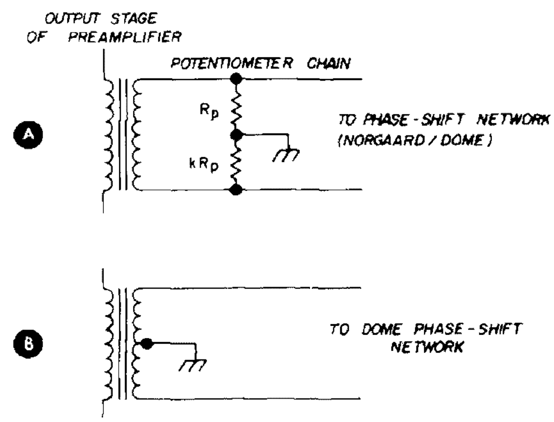



**fig. 9. Usual method of coupling audio preamp to either network, A. Center-tapped transformer should be used for the Dome network, B.**

## incorrect drive ratio

As mentioned previously, the correct drive ratios for the Norgaard and Dome networks are 2/7 and 1. Even very small deviations from these values will result in degraded performance **(fig. 8)**. Allowing for other imperfections in the system, the drive ratio for the Norgaard network

should be kept within ±2 percent, or within the range 0.28 to 0.29. The Dome network is similarly affected, but to a greater extent. In this case the maximum tolerable error is about 1 percent of optimum.

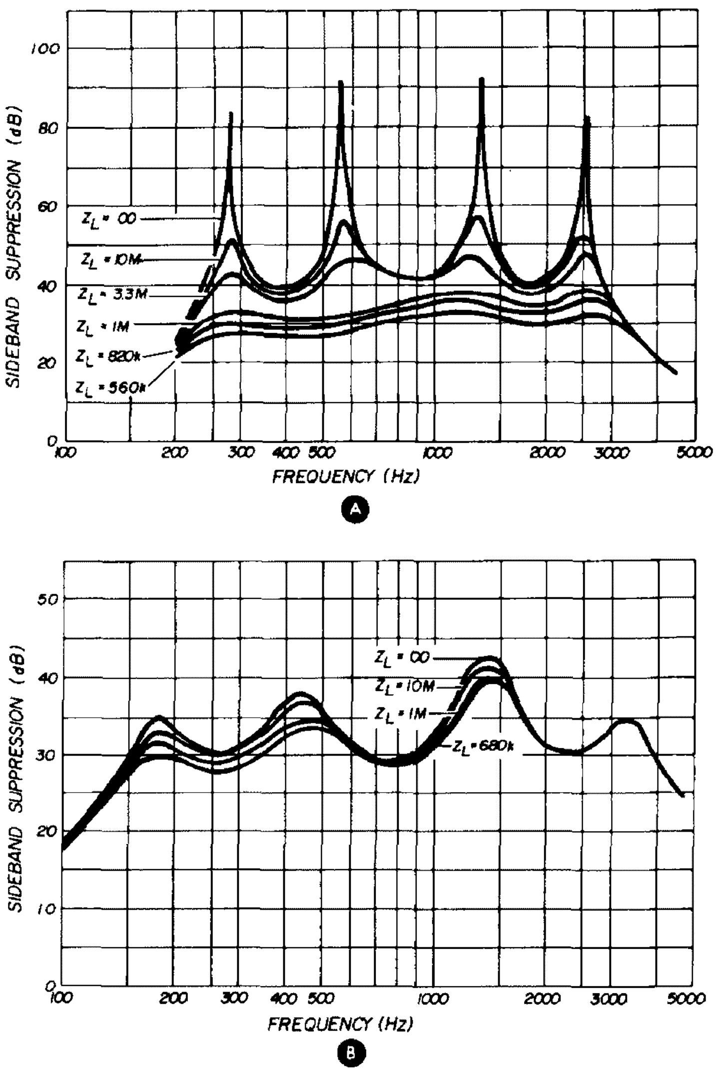



**fig. 10. Effect of terminating outputs of Norgaard network, A, and Dome network, B, in a finite impedance.**

## practical considerations

Input signals to the network are usually transformer coupled from the audio preamplifier. Amplitude ratio is obtained by resistive divider **(fig. 9A)**.

The voltage ratio of the phase-shift networks is extremely critical. Common practice is to include a preset potentiometer across the transformer secondary to adjust this ratio. A potentiometer just doesn't have the long-term stability to maintain a *fixed* and *accurate* ratio of the two input signals. The tapped resistor network should be considered as part of the phase-network. The same care should be used in choosing its components as those of the phase-shift network.

The actual values of the resistors in **fig. 9A** are not important provided they are in the correct ratio. They should be comparatively low in value and consistent with drive capability and terminating resistance of the audio preamp. This will minimize unwanted phase shift due to stray capacitance.

Regarding the Dome network, the resistive divider should not be replaced with a center-tapped transformer **(fig. 9B)**, because the voltage amplitudes across the two halves of the winding won't be within 1 percent of each other over the passband.

## terminating impedance

So far I've discussed network performance under ideal conditions (infinite terminating impedance). It is instructive to consider performance when the networks are terminated in a finite impedance.

The Norgaard and Dome circuits are compared in **fig. 10**, in which each is terminated in an infinite impedance and several values of finite impedance.

A 10-megohm load doesn't seriously affect sideband suppression of the Norgaard network (fig. 10A). However, a 1-megohm termination makes quite a difference. For even lower values, the curves flatten at the low-frequency end and are shifted further downward. Suppression is less than 30 dB over much of the passband with load impedances below 500k.

Load constraints on the Dome network are even more critical **(fig. 10B)**. Terminating impedance for this configuration should be of the order of 10 megohms.

## transistor applications

With both networks, minimum acceptable terminating impedance is very high compared to the input impedance of

conventional transistor amplifiers. Probably fet's should be used here. It might be possible to replace output resistors R5 and R6 (Dome network) with transistor amplifiers that have an input impedance equal to the resistors. Care must be taken in amplifier design to ensure constant imput impedance.

## capacitive loading

When the Norgaard circuit is used with tubes, resistors R2 and R4 (**fig. 4A**) also function as grid-leak resistors, this eliminates any additional loading. Stray capacitance across the network output must be considered in this application, however.

The solid curve of **fig. 11** shows the suppression of the Norgaard network when each output is terminated in a complex impedance, consisting of a parallel combination of 20 pF and 10 megohms. Such a small value of capacitance has negligible effect at the low-frequency end of the passband; but as the frequency is increased, the suppression peaks are lower and shift slightly lower in frequency. Performance is still satisfactory, however, because the suppression doesn't fall below 38 dB except at the extreme high-frequency end of the passband.

The Dome network also may be used with the output resistors acting as grid-leak resistors, but stray capacitance will have negligible effect because the outputs incorporate large shunt capacitors (C5, C6 in **fig. 4B**).

## audio preamp frequency response

When followed by the Norgaard network the preamplifier's frequency response should be as shown in **fig. 12**. From 300 to 3000 Hz the response is flat as is taken as the reference level of zero dB. The suppression of the unwanted sideband over this range is always better than 30 dB, and the network's amplitude response is quite flat (**fig. 7**). Below 200 and above 3000 Hz, preamplifier frequency response follows the curve of **fig. 6**.

Frequency response of the preamplifier followed by the Dome network is similar, except that the curve dips slightly at mid-band. The curve of **fig. 12** shows the overall response of the preamplifier. This is determined by coupling capacitor values and transformer response, in addition to the contribution of a filter. The response curve's slope at the extreme ends of the passband is about 15 dB/octave.

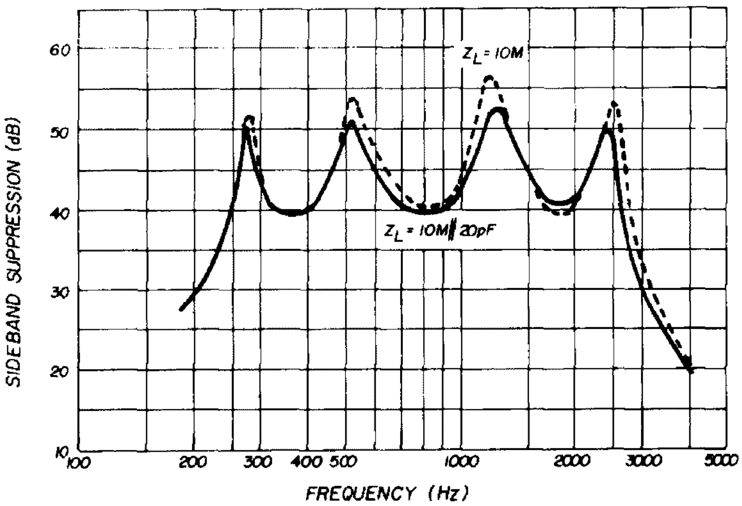



**fig. 11. Performance degradation due to stray capacitance in Norgaard network. Large capacitances shunting Dome networks eliminate this problem.**

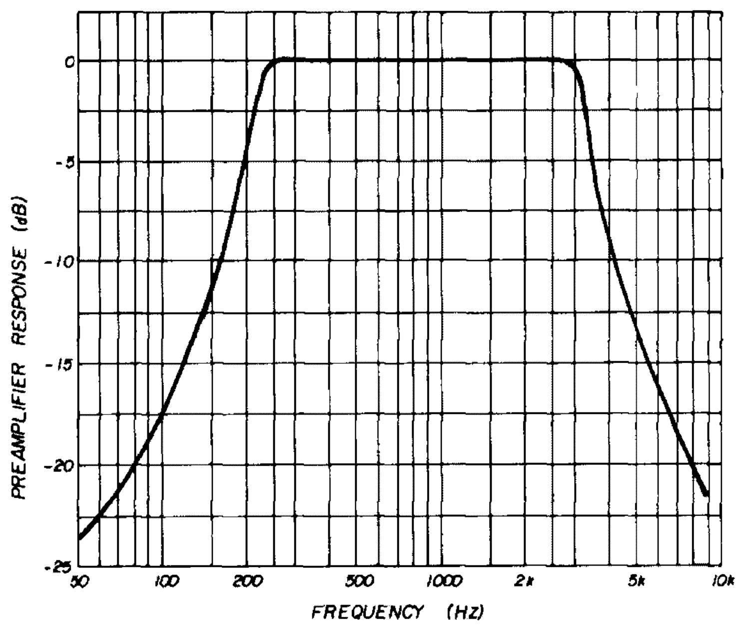



**fig. 12. Audio preamplifier frequency response. A slope of about 15 dB/octave is required at each end of the passband to ensure 30-dB sideband suppression.**

**ham radio**

## keyer oscillators

Two oscillators are used in the electronic keyer mentioned above. One, the clock, forms the dots and dashes; the other is a sidetone generator. Both oscillators are of the form shown in fig. 1. Each uses three-quarters of a quad two-input NOR gate, the LU380A. The oscillator consists of three NOR gates arranged in a ring. The only other components in the basic circuit are a timing resistor and capacitor. The rings consists of an odd number of inverting gates; thus it has no stable state and tends to oscillate at a frequency determined by C1 and R1. The frequency is roughly equal to 1/(2RC), where frequency, R, and C are in Hz, ohms and farads. The maximum obtainable frequency is about 5 MHz. The output is a pretty fair square wave at the lower frequencies.

## clock oscillator

For the clock oscillator, we use the circuit shown in **fig. 2**. The keyer is designed to operate at 8 wpm for the lower end of the speed range. Using readily available components, and trying to keep the value of R2 + R3 fairly low for linearity of adjustment, we find that a value of 25 µF for C2 and C3, 5k for R2, and 1k for R3 will yield a frequency of 3.3 Hz. Even though C2 and C3 are in series, they effectively yield their individual capacitance values. It turns out that 3.3 Hz works out to 7.92 wpm. This is based on:

Code speed (wpm) = 2.4 x dot frequency in Hz

The origin of this formula is obscure, but it correlates closely with the other rule-of-thumb, which is:

Code speed (wpm) = no. of dashes in 5 seconds

Tests show that the calculated value is very close to that measured.

To find the upper end of the speed range when R2 is cranked down to zero resistance, we find the theoretical upper speed is 47.52 wpm. This, again, agrees closely with the value found in practice. In the actual keyer, R3 is selected to meet the 50 wpm mark exactly.

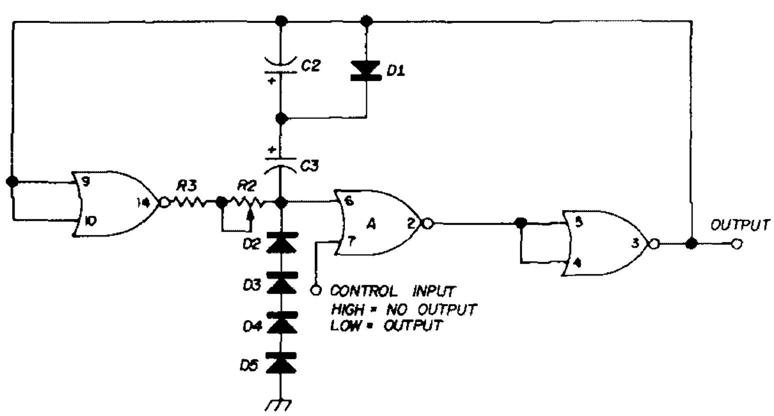


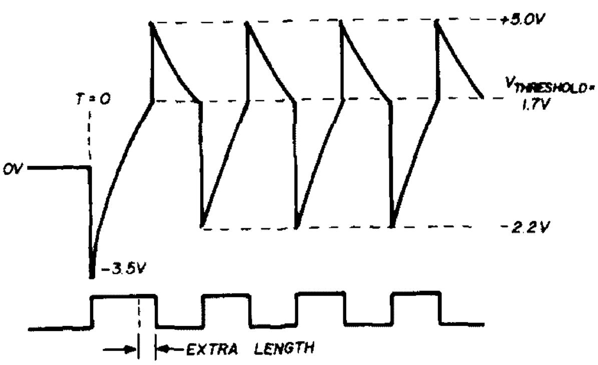

**fig. 2. Clock oscillator circuit.**

Utilogic is a registered trademark of the Signetics Corporation.

## operation

Diode D1 shunts reverse bias around

**fig. 3. Timing diagram showing effect of eliminating gate input clamp.**



C2 when the keyer is idling. Otherwise, the capacitance of C2 would slowly disappear due to its reverse polarity. The best choice for the capacitor would be a nonpolarized type, but it would be as large as the whole keyer. C3 needs no protection, since it has a very low duty cycle of reverse bias.

Diodes D2 through D5 are a negative clamp on the gate input to ensure that the first cycle of the oscillator is close to the same length of succeeding cycles. **Fig. 3** shows the clock waveform without these diodes. The first cycle starts with the input of gate A resting near ground potential, but subsequent cycles start at the threshold point of the gate. The error, in any case, is not great and can be discerned only with instruments; and the resulting cw sounds fine either way.

**Fig.** 4 shows the error after correction. It also shows the high degree of dot/space ratio accuracy obtainable with this simple circuit.

The control input of gate A inhibits oscillation when it is held at logic state 1.

tor frequency, normally 800 Hz, is adjustable by R4. The output is gated by NOR gate B, which drives the base of a 2N404 through current-limiting resistor R5. An

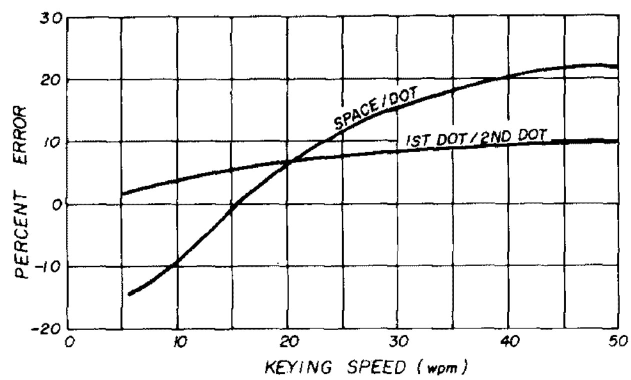



**fig. 4. Timing accuracy of the clock circuit after correction. Note high degree of dot/space ratio.**

ordinary output transformer, with a 500-ohm center tapped primary and 8-ohm secondary driving a 2-inch speaker, completes this simple monitor circuit.

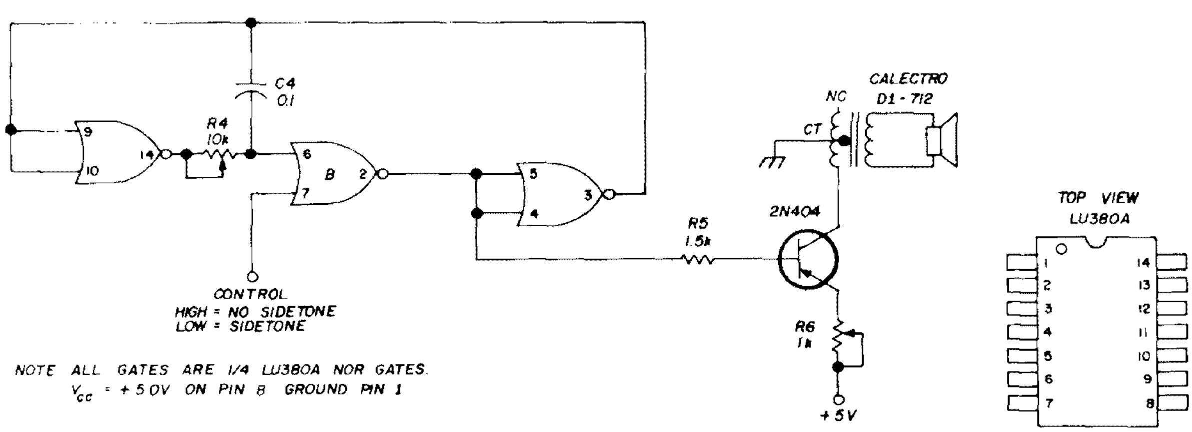



**fig. 5. The sidetone oscillator. Frequency is adjustable by R4; output level by R6.**

When the control input is lowered to logic state zero, the oscillator starts generating the first character. I found this instant-starting clock much simpler to use than the free-running variety, which tends to force spacing between letters.

## sidetone oscillator

The sidetone oscillator, **fig. 5**, is of the same form as the clock. Sidetone oscilla-

The output level, controlled by R6, is adequate for a roomful of people. Fairly high efficiency is obtained, since the 2N404 operates near class C.

### reference


1. L. Brock, "Utilogic NOR and OR gate Applications," Signetics Corporation Applications Memo 97, June, 1969.


**ham radio**

THE HAMMARLUND MFG. COMPANY
PERFORMANCE DATA FORM
LINEAR AMPLIFIER HXL-1

SERIAL NO. 33840268

DATE 3-11-70
BY JP / C. [illegible]

20 Ma. IDLE PLATE CURRENT (RELAY OPEN)
100 Ma. IDLE PLATE CURRENT (RELAY CLOSED)

NEUTRALIZATION

80M ✓ 40M ✓ 20M ✓ 15M ✓ 10M ✓

2500 VOLTS PLATE VOLTAGE ON STANDBY

PREPARED BY:

ROBERT MILLER, W4AEY
ENGINEERING LEADER

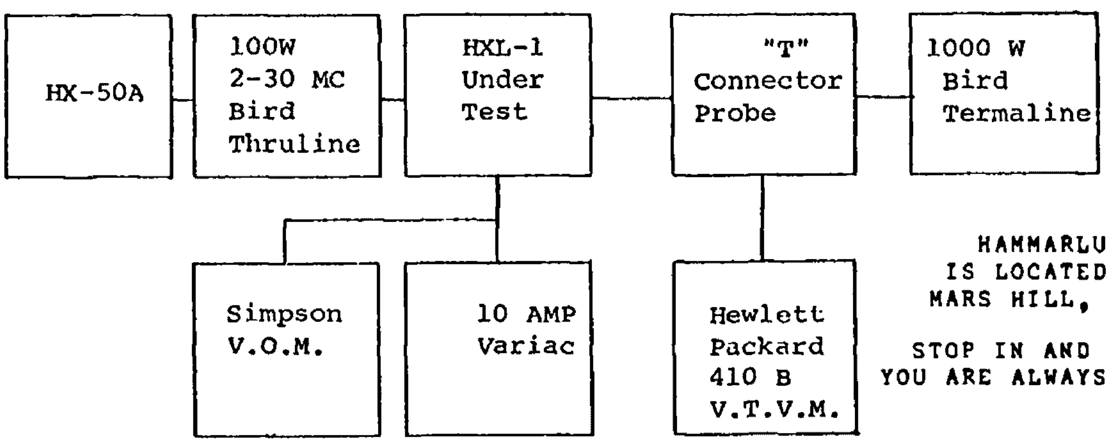
USE USB 2 TONES OF EQUAL AMPLITUDE

NOTE: ADJUST LINE VOLTAGE TO 110 VOLTS UNDER LOAD FOR POWER MEASUREMENTS.

| BAND FREQUENCY | RF DRIVE * * WATTS | PLATE VOLTAGE * * VOLTS | PLATE CURRENT * * MA | POWER OUTPUT * * WATTS | % EFFICIENCY * * | P.E.P. OUTPUT * * WATTS |
|---|---|---|---|---|---|---|
| 80M 3.8 MC | 100 | 1800 | 650 | 860 | 73.5 | |
| 40M 7.3MC | 100 | 1725 | 600 | 840 | 81.0 | |
| 20M 14.3MC | 100 | 1700 | 608 | 840 | 81.0 | |
| 15M 21.3MC | 100 | 1600 | 725 | 810 | 70.0 | |
| 10M 28.3MC | 95 | 1650 | 900 | 820 | 55.0 | |

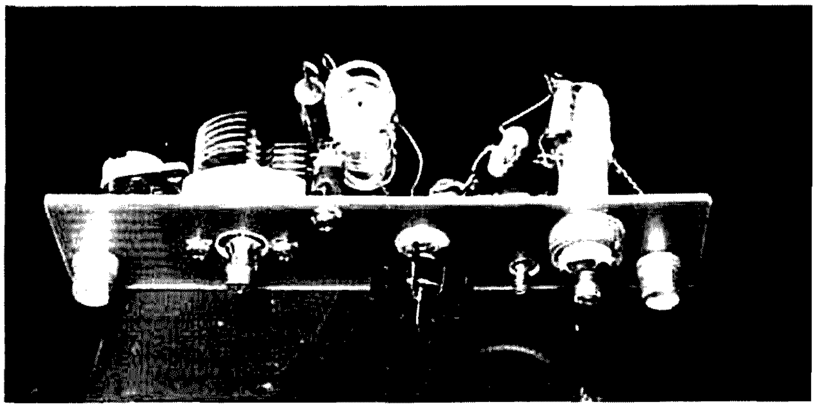
EQUIPMENT SET-UP

MULTIPLY x 2.5 FOR PEP  HELP YOURSELF!




HAMMARLUND
IS LOCATED IN
MARS HILL, N. C.

STOP IN AND VISIT.
YOU ARE ALWAYS WELCOME.



# transistor frequency multipliers

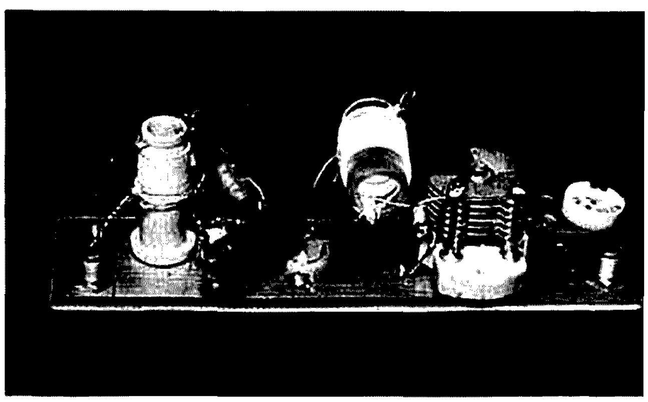
Selecting transistor frequency-multiplier circuits for maximum efficiency and output

Frank C. Jones, W6AJF, 850 Donner Avenue, Sonoma, California 95476

The circuit board in the photograph was put together to try out different circuits for frequency doubling from 14 to 28 MHz; a high-output signal generator provided the 14-MHz drive. A wattmeter across the output provided readings for comparing the relative efficiencies of different multiplier circuits.

The transistor used for nearly all the tests, the 2N5188, is an inexpensive npn transistor rated at 4 watts maximum dissipation at 25° C; this indicates a maximum dissipation of 2 watts or so with a moderately sized heat radiator slipped over the transistor case. Typical cutoff frequency for the 2N5188 is 325 MHz. Selected 2N5188's can be used in class-C rf service up to about 150 MHz—at lower frequencies nearly all units give equally good results without special selection.

An rf voltmeter was connected from base to ground (or emitter) to make sure the maximum base-to-emitter breakdown voltage of 5 volts peak was not exceeded during tests. The proper value for the base-bias resistor in frequency multipliers ranges from 100 to 5000 ohms. If the base-bias resistance is too great the tran-



sistor may be damaged by high rf drive since the 5-volt rating is easily exceeded. In most of these tests a 1000-ohm bias resistor was used, and the rf drive adjusted to 3 volts rms maximum.

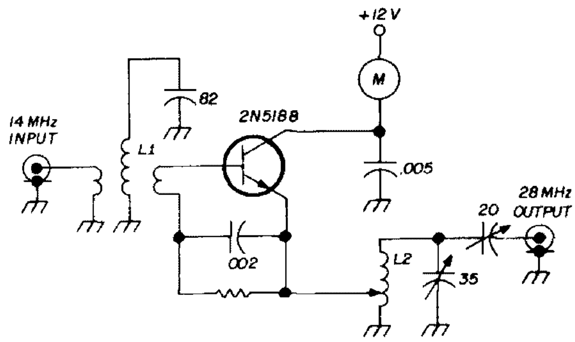



fig. 1. This circuit is essentially the same as fig. 3 although the transistor collector is at rf ground.

The 2N706, 2N2711 and 2N222 provide moderately good efficiencies when operated as frequency multipliers up to about 1/5 of their cutoff frequency. The 2N4427, 2N3866, 2N3553 and similar power transistors provide more output on two meters than the 2N5188, but they cost several times as much. Two 2N5188 stages—one as a frequency multiplier providing 0.1 to 0.3 watts output and the other as a buffer with 0.5 to 1 watt output—offer an economical way of building a two-meter transmitter. The more expensive power transistors can then be used as power amplifier stages for a few watts of fm or cw.

## practical circuits

Four practical frequency-doubler circuits are shown in fig. 1 to 4. Although the circuit of fig. 3 is the one usually seen in the literature, the circuits of fig. 2 and fig. 4 proved to be far superior. The circuit of fig. 1 is a little unusual in that its collector is grounded for rf, but it is basically the same as fig. 3 (and actually performed about the same). Fig. 1 has slightly less rf radiation loss since the collector is connected to the case—the construction used for most transistors that are rated at more than 0.5 or so. However, in the 14-to-28 MHz frequency doubler tests the rf loss wasn't high enough to warrant the use of the common-oscillator layout over the more usual grounded- or common-emitter circuit.

The reason the circuits in fig. 2 and fig. 4 gave twice as much output as the others is quite simple: the transistor has a relatively high feedback capacitance between collector and base, so any appreciable impedance from base to emitter—even a single-turn link—causes degeneration; the series-tuned 28-MHz circuit in fig. 2 provides a very low impedance path from base to emitter and effectively connects the collector to-base feedback capacitance across the output. This eliminates degeneration and increases output by 1.5 to 3 times in doubler, tripler and quadrupler stages.

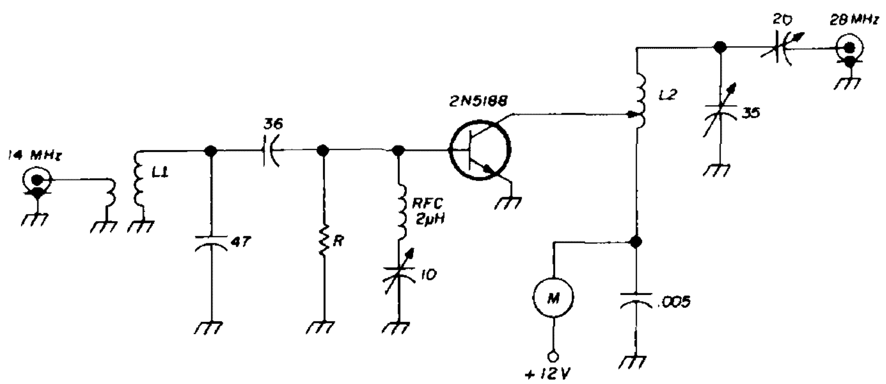



fig. 2. This circuit provides a low-impedance path from base to emitter and effectively connects the collector-to-base feedback capacitance across the output.

The series-trap circuit also provides excellent efficiencies at vhf and uhf but it requires careful adjustment. The disadvantage of the circuit in fig. 2 is that it will break into self-oscillation if the series-

resonant circuit is incorrectly tuned; it also requires an additional adjustment in the frequency multiplier.

The advantage of the series-tuned circuit is that it can be applied to an existing frequency multiplier in a vhf or uhf converter or transmitter to obtain more output. Usually a small moderately high-Q rf choke coil can be selected which will resonate with a 5- or 10-pF piston capacitor to the desired *output* frequency. With this simple modification tripler and quadrupler stages really come to life as far as rf output is concerned.

The frequency-multiplier circuit in **fig. 4** is my favorite for doublers, triplers, or quadruplers—at high or low power. It provides a low-impedance path from base to emitter if C2 is large enough to provide a reactance of only a few ohms at the output frequency; if the reactance is below 10 ohms excellent results can be obtained. The ratio of C2 to C1 should be between 5:1 and 10:1.

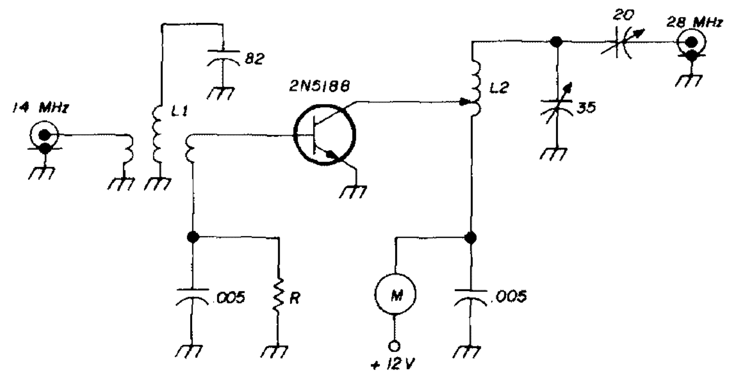



**fig. 3. This is the frequency-multiplication circuit usually seen in the literature.**

In most small-power frequency-multiplier circuits the collector must be connected to a tap on the output tuned circuit to obtain a reasonable impedance match. To minimize undesired harmonic output the tuned circuit should have an operating Q of at least 15. In usual designs the collector tap is located 1/6 to 1/2 the total number of turns from the cold end.

Low-power frequency-multiplier circuits, such as those used in vhf receiving converters, can be designed with the collector connected to the hot end of the tuned circuit if the tuned circuit has a resonant impedance in the range of 1000 to 3000 ohms. If the circuit doesn't tune sharply it indicates low circuit Q and poor harmonic suppression, and may mean the collector load or external load is too heavy.

**ham radio**

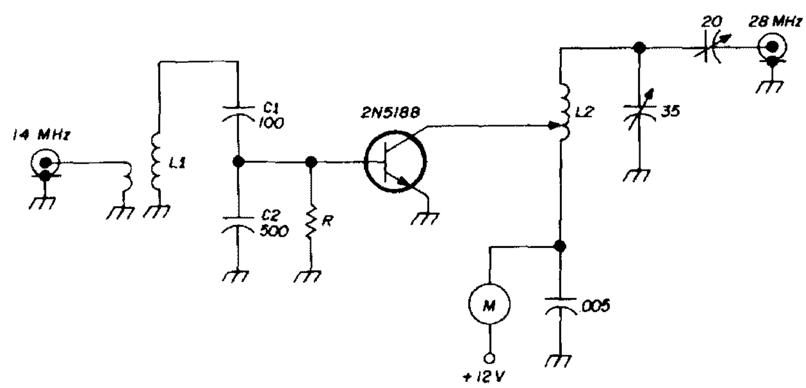



**fig. 4. This circuit is W6AJF's favorite for frequency-multiplication service.**

# rtty frequency-shift meter

This instrument will measure rtty shift as well as small frequency differences in any audio filter

Jim Goding, VK3ZNV, 15 Yarrabee Court, Mount Waverly, Victoria, Australia

One of the problems confronting the rtty operator is measuring the correct frequency shift between mark and space signals.

This article describes a direct-reading indicator with which you can measure these frequencies from your receiver output. By offsetting the mark frequency to zero and increasing meter sensitivity, the difference between mark (2125 Hz) and space (2295 or 2975 Hz) can be accurately measured.

## the circuit

The circuit consists of an input amplifier, Schmitt trigger, monostable multivibrator, averaging amplifier, and meter amplifier (fig. 1). The input amplifier has an input sensitivity of about 100 mV. The Schmitt trigger squares the input signal, which is then differentiated by the 300-pF capacitor. The negative pulses are clipped by a diode, and the positive pulses trigger the monostable (one-shot). Thus a series of rectangular pulses is produced whose repetition rate is equal to the signal frequency. Pulse amplitude and length are respectively 3.6 V and 80 microseconds.

The averaging amplifier produces a steady negative dc voltage proportional to the average voltage of the waveform at the one-shot's output which is, in turn,

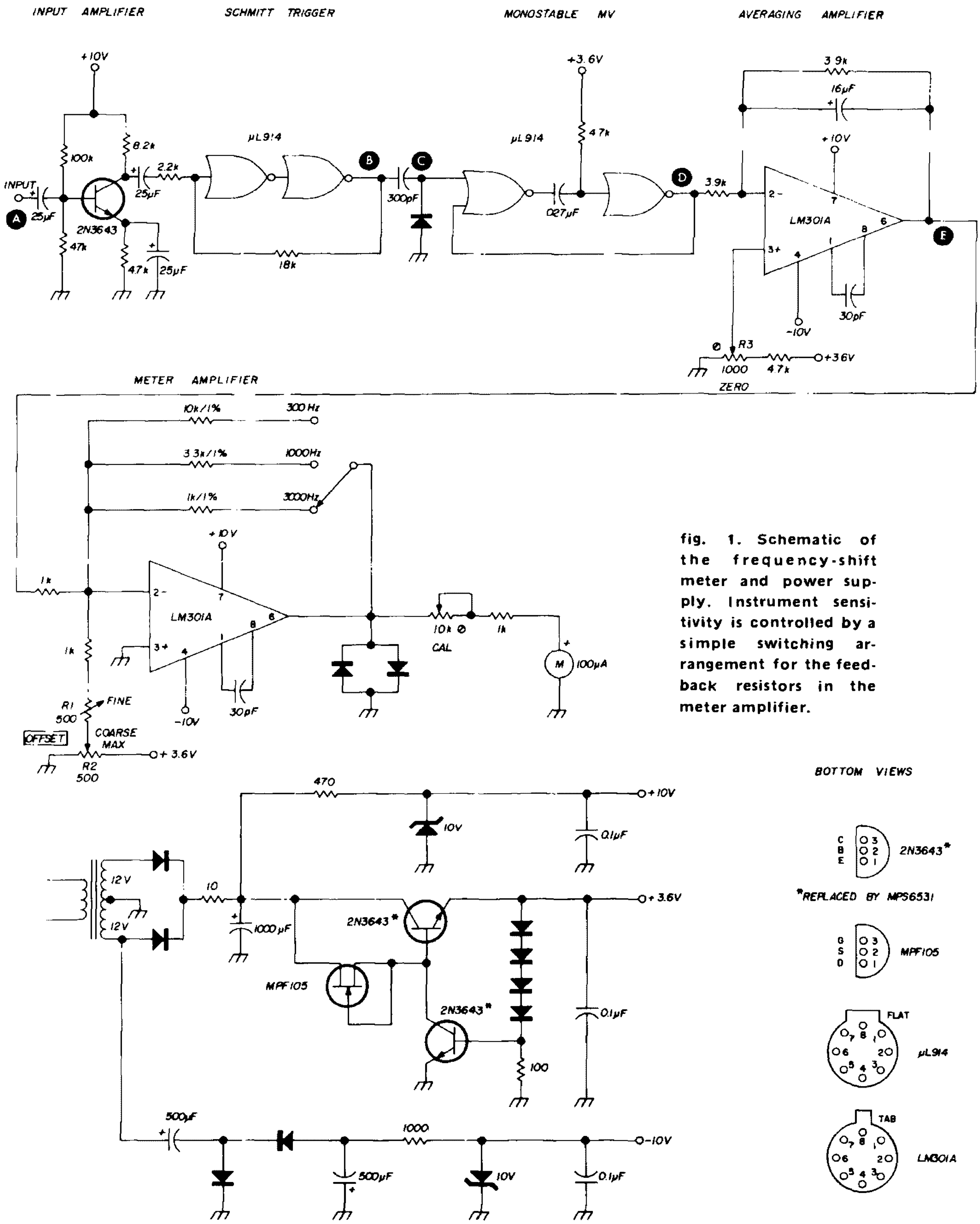



fig. 1. Schematic of the frequency-shift meter and power supply. Instrument sensitivity is controlled by a simple switching arrangement for the feedback resistors in the meter amplifier.

proportional to input frequency. Typical waveforms are shown in **fig. 2.**

Instrument sensitivity is selected by switching the feedback resistors in the meter amplifier. The meter amplifier will have a gain of 1, 3.3 or 10 corresponding to 3000, 1000 and 300 Hz respectively. Potentiometers R1 and R2 allow a small current opposite in polarity to the signal current to be applied to the meter ampli-

fier; thus the meter indication is fully adjustable.

## calibration

An accurately known audio frequency is required for calibration. With no input to the instrument, set offset controls R1 and R2 to zero and the range switch to 300 Hz. Adjust R3 for a zero indication on the meter. Set the range switch to the appropriate range for the calibrating frequency; i. e., the range that gives the highest reading without pinning the meter. Next apply the calibrated signal, and adjust the calibration control for the correct meter indication.

If 1 percent resistors are used as shown, only one range need be calibrated; the others will be automatically correct. Zero-set and calibration controls may be preset screwdriver-adjustment pots located on the back of the instrument.

## shift measurement

To use the instrument, set the range switch to 3000 Hz and apply the mark frequency to the input. The meter should indicate the mark frequency. Set the meter to zero, *with the mark frequency still applied to the input.* Then switch the meter to the 1000-Hz range for 850-Hz shift, or to the 300-Hz range for 170-Hz shift. The zero point might need readjustment with the fine offset control. Now apply the space frequency to the input. The shift may then be read on the meter.

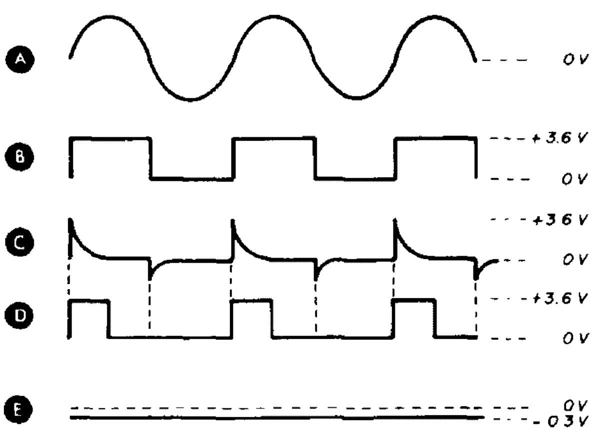



**fig. 2. Typical waveforms. These will appear at similarly labeled points in the schematic.**

## meter protection

While the circuit is virtually foolproof, it's possible to pin the meter in the forward direction if the input signal's frequency is too high. The meter can be pinned in the reverse direction when no input is applied with the offset control advanced. Therefore, the diodes in the meter amplifier output are included to protect the meter from possible high overload levels.

## noise response

If noise is on the signal to be measured, as when measuring the shift of a weak signal, errors may result because the Schmitt trigger will fire on noise pulses as well as on signal. The error may be reduced by using the minimum audio signal that will give a steady reading. Errors due to noise will be no problem, of course, when setting your transmitter's shift by monitoring its signal on your receiver.

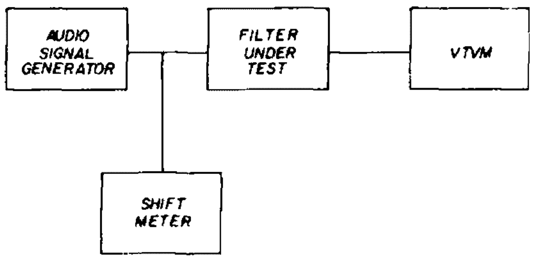


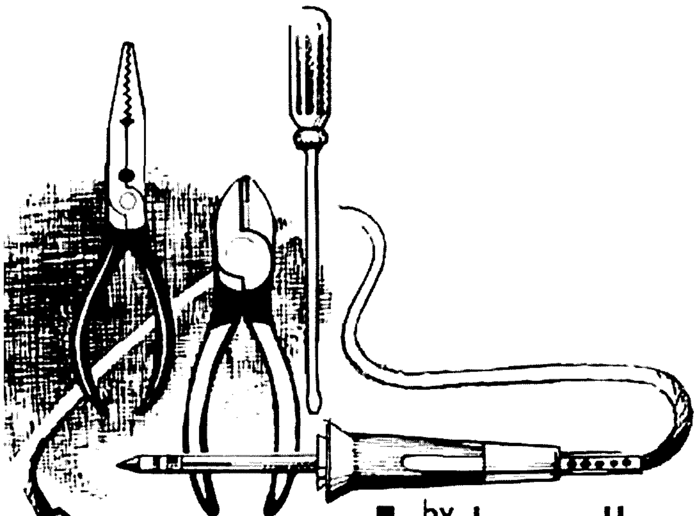

**fig. 3. Application of the circuit for adjusting small differences in audio filter output frequency.**

## other uses

The instrument is also useful in adjusting audio filters for rtty and other uses. A typical test setup is shown in **fig. 3.**

While most audio signal geneators have good *absolute* accuracy, their calibration is seldom good enough for measuring small differences in output frequency. In setting up many filters, particularly for rtty, filter bandwidth is often more important than center frequency.

**ham radio**



by larry allen

# the repair bench

## curing trouble in mobile power supplies

Remember the days when only three things could go wrong with a mobile power supply? A sticky vibrator, a bad buffer capacitor, and—if things were really bad—a transformer.

It wasn't hard to figure out which it was; they all blew the fuse. A done-in transformer spewed was and varnish, and a worn-out buffer made even a new vibrator sound ragged. And substituting a new vibrator was a sure way to bring the output voltage back up to normal if the old vibrator was shot.

Nowadays when a fuse pops, it all happens so quietly you hardly think anything's wrong at all. The buzzing vibrator has been replaced by transistor switches. The frequency they flip-flop at is audible but so quiet you can't usually hear it. And when one of the transistors goes—it generally goes completely. There's no bounce, rattle, or noise. Just thump! and the fuse or circuit breaker pops open.

## the heavy demands

A lot is expected of the mobile power supply. Take the one pictured in fig. 1. It's big enough to operate an ssb linear from a 12-volt car battery. It can furnish up to 500 mA at 2 kV, if it has to. Typically, it runs about 2.1 kV with an average load of 180 mA. It also supplies -110 volts at 60 mA, for bias.

The switching transistors in a supply like this must be heavy ones. Likewise the transformer and rectifier diodes must be rated to withstand lots of voltage and current. Input wiring and components must carry the heavy battery-current drain of a supply like this—as much as 30 or 35 amps.

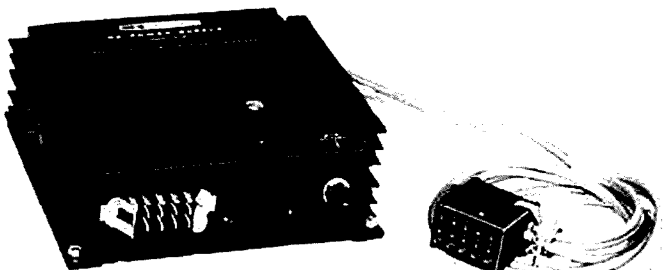
Running a unit like this on the repair bench requires the same kind of care you'd use installing it. Heavy wiring from the dc supply is a must. No ordinary battery eliminator today can handle the load, so you'll need a storage battery. A charger across it will keep it at full power.

An ideal bench setup brings the 12-volt battery cables up to ¼-inch bolt terminals at the back of the bench. A 4-foot cable is made up from two No. 6 awg wires terminated at one end in soldered-on eye terminal lugs; the holes in the lugs fit over the terminal bolts. The other end of the test cable has a heavy-duty Jones female plug to fit this power supply. Other cables can be made up for other power supplies.

A 60-amp cartridge fuse (Buss FRN-60) in the battery-cable line protects the battery. It doesn't take long for a dead short to ruin battery plates and burn the insulation right off the cable wires. If you set up your bench like this, occasionally spin the cartridge fuse in its holder; that keeps the contact clean and cool.

## the switched inverter

The basic input circuit of the Heath



**Heathkit HP-13 dc supply, their newest model, is designed for mobile operation of the SB-101, SB-110A, HW-100 and Heath Single Banders.**

mobile supply is diagramed in fig. 2. A 40-amp breaker protects the car battery. An input relay is necessary because no ordinary switch could carry the power for this supply. A switch on the transmitter closes the relay. A choke and capacitor decouple the input dc line, so no hash

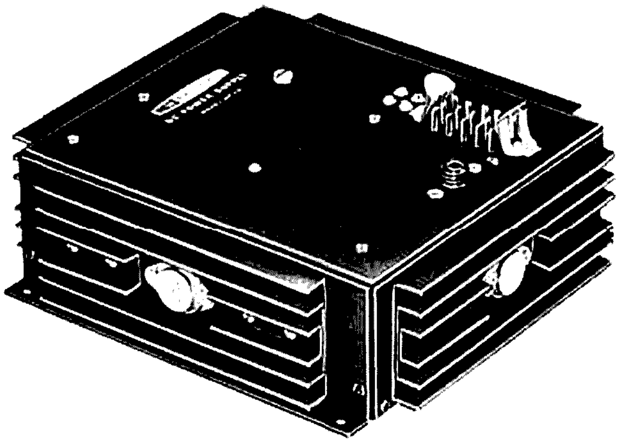


**fig. 1. Heathkit HP-14 power supply, although no longer marketed, is an example of a high-voltage mobile power supply that can be used with a 1-kW PEP linear amplifier. Switching transistors are heavy-duty types, and heavily heat sinked. (Two are out of sight on other sides of the unit.)**

from the dc-to-ac inverter can affect other accessories.

When the input relay closes, 12 volts positive is applied to the emitters of both transistors through a short portion of the transformer winding. The collectors are grounded, making them negative with respect to emitter. That's the first requirement for operation of pnp transistors.

At the same time, the rest of each winding half is feeding the positive voltage to the bases and to resistors R1 and R2. The voltage at the base of each transistor is less positive (thus more negative) than that reaching the emitter, because the winding drops some of it (R1 and R2 are dividers with the winding resistances). With base negative (to the emitter) the pnp transistors can conduct.

However, the transistors are not perfect matches. One conducts more than the other. Heavy electron current flows in one (for this example, let's say it's Q1). The path is from ground through collector to emitter, through a short portion of the transformer winding to the center tap, and out to the positive battery terminal.

As this current builds up, it develops a magnetic field in the winding. That field applies a negative-going rise to the base of Q1, adding to the negative bias already there. The transistor conducts even more.

Meanwhile, the same magnetic field is backward-biasing the other transistor. The portion of transformer winding between Q2's base and emitter applies a

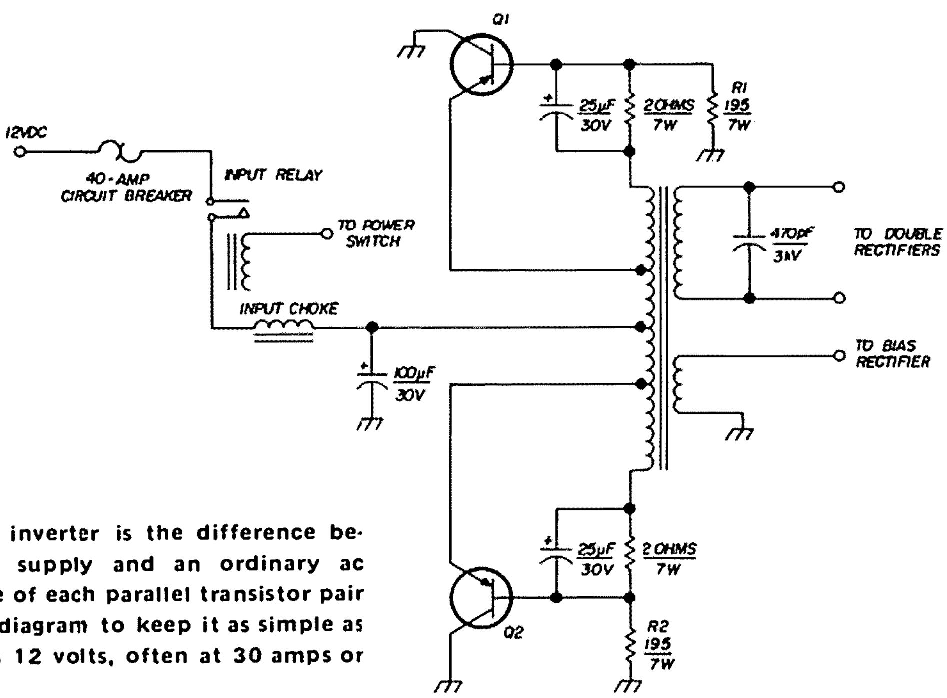



**fig. 2. Dc-to-ac inverter is the difference between a mobile supply and an ordinary ac supply. Only one of each parallel transistor pair is shown in this diagram to keep it as simple as possible. Input is 12 volts, often at 30 amps or more under load.**

positive-going voltage to the base, which cuts that transistor off. No current can flow in the collector/emitter/winding circuit of that transistor.

Sooner or later, the transformer saturates. That is, even though current in Q1 keeps increasing, the magnetic field doesn't. Then, no more extra base bias for Q1 is developed by the transformer winding. Q1 current suddenly begins dropping off.

As the magnetic field collapses, the whole process reverses itself. Suddenly, Q2 gets the go-ahead bias it needs to start conducting. And it does. The more it does, the more its end of the transformer winding forward-biases it. It takes off like Q1 did earlier. Current goes way up. And all this time, Q1 is being cut off like Q2 was earlier.

This goes on till the transformer saturates at the Q2 end. Then again the process reverses and Q1 takes over. The transistors keep switching back and forth, generating more or less of a square wave. There's a photograph of how it looks on the oscilloscope in **fig. 3**. Notice how transformer saturation tapers the voltage off gradually until suddenly the other transistor takes over. Then the waveform has straight sides till full voltage is reached in the other polarity.

Another factor that rounds the trailing corner of the output waveform (and causes the overshoot at the leading corner) is the capacitor across the high-voltage secondary. Without it, transient overshoots and preshoots could generate large counter-emf's that might damage the transistors. In this respect, the 470-pF capacitor acts like the buffer capacitor in old vibrator supplies.

A resistor-capacitor network in the base circuit of each transistor is further protection; they absorb any transient spikes that might zap the base junctions of the transistors.

## trouble in the switcher

The most likely trouble is a shorted transistor. As mentioned on the schematic, each transistor is actually a pair; there are four transistors, with heat sinks. If a transistor shorts, the switcher quits. Usually, the breaker opens—from the short or from overcurrent in the two opposite transistors.

To test the shorted one, you have to disconnect the base and emitter leads. Usually, the leads have tips that slip down over the pins. If the wires are soldered, don't overheat the pins disconnecting and reconnecting them.

Use your ohmmeter. Clip the common test lead to the collector (grounded in

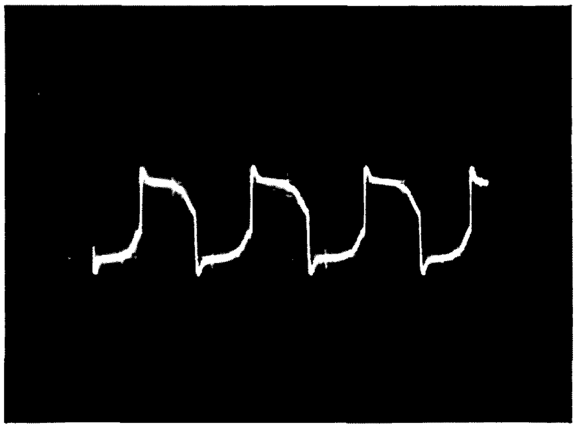


**fig. 3. Waveform developed by dc-to-ac inverter. This shaped square wave is what is fed to the rectifier doubler circuit by the transformer secondary.**

this example). The ohmmeter should read open or extremely high to both pins. Next clip the ohmmeter common to the base pin and touch the probe to the emitter. Then reverse them. You should get a low reading (but well above zero) one way and a high reading the other. If you get an infinite reading both ways, the transistor is open. If you get zero or a very low reading both ways, the transistor is shorted.

Be cautious replacing the transistor. Be sure it's the right type. If it opens, the others overload and may go. Position the new one carefully. Sometimes the mounting holes are sloppy and the base or emitter can touch chassis ground. If it's the base, goodbye new transistor!

If a transistor blows, also check the base resistors and electrolytic capacitors (with one end loose). It only takes a few

minutes. A poor electrolytic may let transmitter keying transients damage the transistors. If so, a new transistor may soon go the way of the old one.

If you have a capacitor checker with a power-factor test, use that. Also measure capacitance. If not, use your ohmmeter. Make sure both electrolytics charge up quickly and don't show a resistance below 50k. If they charge slowly, power factor is too high. You can measure resistors directly.

charges across the two strings of capacitors add, and the output is double the input.

The bleeders, R1 through R7, have a number of purposes. For one thing, they're a safety factor. They drain off any charge on the capacitors, and on any capacitor in the transmitter, when the supply is shut down. Also, connected as they are, they equalize voltage distribution across all the electrolytic capacitors. Without them, one capacitor might devel-

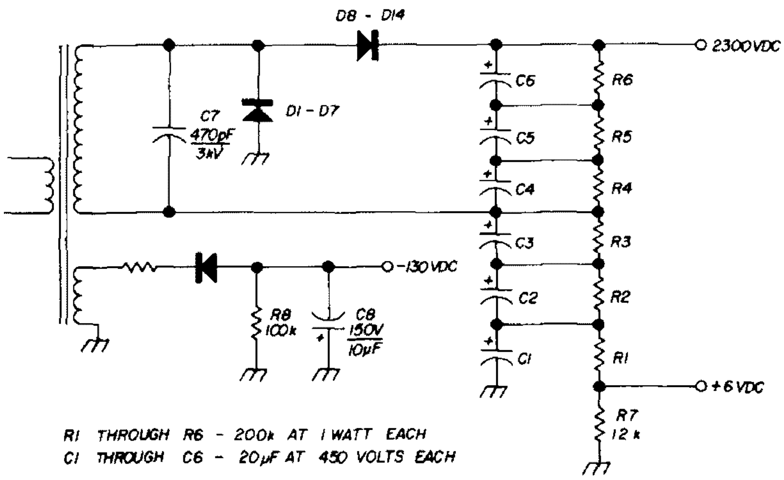



**fig. 4. Dc output section of typical mobile hv supply. Diodes in series increase voltage rating without expense of hv diodes. Circuit is full-wave doubler. Bleeders distribute voltage among capacitors, which are also in series for higher voltage rating.**

## the high dc side

You can have trouble with the dc circuits of a mobile supply, too. That consists of rectifier diodes, filter capacitors, and bleeder resistors.

The output of a supply in this example is diagramed in **fig. 4**. The transformer is the one you see in **fig. 2**, but only the secondary is drawn here.

The transformer steps up the voltage of the primary waveform, which is only about 10 volts peak to peak (p-p) at the emitter of the transistors. There's about 1000 volts p-p across the secondary winding and C7.

Diodes in series give a rating to withstand such high voltage. There are seven diodes in each leg of the voltage-doubler circuit.

The doubler is a simple full-wave type. On one half-cycle, diodes D8–D14 charge capacitors C4–C6. One the next, diodes D1–D7 charge capacitors C1–C3. The

op too much voltage and break down.

Also, R7 is a small-value divider resistor in series with the six main bleeders. It develops a 6-volt positive dc. This output is ordinarily for automatic level control (ALC) bias.

The bottom winding develops about 125 volts p-p. That's rectified in a diode connected with anode to output. A filter capacitor and bleeder resistor keep the output at a negative 130 volts dc (-130 V dc). That's bias for the linear amp. In one Heath linear, it also operates the antenna relay.

## dc output troubles

Some troubles in a supply like this can be diagnosed from the symptoms.

Low output, for example, can be the result of an open diode. That leaves only half the doubler working, and no-load dc output is reduced to half or less.

An open bleeder can cause the parallel capacitor to be damaged. If output drops drastically with even slight load, check each output resistor and capacitor. You may have to disconnect one end of each capacitor to get a meaningful ohmmeter or checker reading.

Whenever you find an open bleeder resistor, always replace its capacitor too. It has taken an overdose of voltage. Even if it seems healed, its ability to form a dielectric may be impaired.

A shorted diode in a series string like this doesn't show up immediately. Later, another may go, then another. Eventually, one will probably burn open. Some may develop backward leakage that will eventually damage others. So, when *any* diode trouble is found check all the others in that series string. Just check them one at a time with your ohmmeter; you don't even have to disconnect them. One direction should read less than 100 ohms (actual amount depends on the ohmmeter battery). The other direction should read open.

It's a good idea to check the bleeders periodically. Just watch a voltmeter connected to the dc output. Turn the supply on, *with no load* connected. Then turn it off. The meter should drop to zero volts in a matter of 1 or 2 seconds. If not, BEWARE. The bleeder is open.

An open R7 shows up in other ways, too. Most notably, the ALC output voltage (6 volts) rises to practically the full dc output voltage. The linear couldn't take that. If the ALC diode in the linear blows, first thing to check is the ALC-bias bleeder in the power supply. An easy was is by the bleeder test just described.

Troubles in the -130 volt bias supply are ordinary. If output is very low under load, R9 or the diode is probably the culprit. If it's low even without a load, the capacitor (C8) is probably bad. Be sure to connect any replacements with proper polarity. Check the 100k bleeder, too. A meter on the output, with no load, should drop to zero as fast as the meter pointer will allow, when you turn the supply off.

## caution—high voltage

You've got to be careful around these supplies. Never assume the bleeder is okay. A jolt of 2000 volts, when current capability is 500 mA, can cool you for good. It only takes 8 or 9 mA to make your heart fibrillate, if you don't get aid within a very few minutes the damage is permanent. Dead permanent. Even a couple of milliamps can hurt an already weak heart, or weaken a good one.

I never work around one of these except with two things: a 1-meg 2-watt carbon resistor I know is good, and a jumper lead. The resistor goes across the output before I turn on the supply the *first time*. It *stays there until* the last test is run and I'm sure the internal bleeders are working. Besides that, I always keep a jumper clipped across the high-voltage output except when I'm firing up the supply for a live test. When I turn it off, back on goes the jumper. This may seem like trouble, but so are funerals.

Whenever possible, I do my troubleshooting with tests that don't require me to fire up the unit. That takes some of the danger away. (I wrote about that kind of troubleshooting in this department in August 1968, page 52; November 1968, page 62, and January 1969, page 52.)

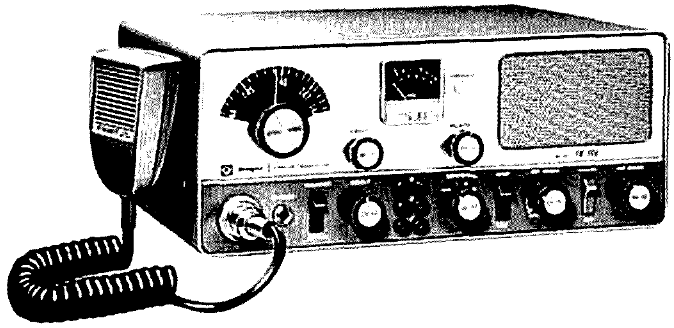
## what's ahead in repair bench

Solid-state equipment is now the rule more than the exception. With it comes the old fear of not being able to keep the gear operating. More specifically, hams worry about how to track down trouble in a solid-state receiver or transmitter.

Most hams can do pretty good deciding what section or stage is fouled up. But their unfamiliarity shows through when they get down to troubleshooting the circuits inside the stages.

Next month, I'll show you a sure-fire approach to servicing transistors. I won't make a super-speed technician out of you, but when you get through reading next month's column you'll know how to find out what's working and what isn't.

**ham radio**



# knight-kit two-meter transceiver

Jim Fisk, W1DTY

If you've been shopping around for a little two-meter rig, you've probably seen the ads for the Knight-kit TR-108 transceiver. This little unit covers the entire two-meter band, and combines a sensitive double conversion receiver with fifteen watts input and a built-in 117-Vac/12-Vdc power supply. All you need to get on the air is an antenna and a crystal for your favorite two-meter frequency (an optional vfo kit is available if you want to move around the band a bit). Also, if you want to go mobile, an accessory mobile mount is available.

## receiver

The first stage of the two-meter receiver section is a 6HA5 grounded-grid rf amplifer. This stage provides adequate gain and noise figure for the maximum range that can be expected with the TR-108. The crystal-controlled oscillator/tripler that feeds the first mixer uses a third-overtone crystal cut to 37.883 MHz—a 6CW4 serves as the oscillator and the triode section of a 6GJ7 as the tripler. The pentode section of the 6GJ7 is used as the first mixer with a resulting output of 30.35 to 34.35 MHz. Since the converter section has a 4-MHz bandpass, each coil must be properly peaked to maintain the broadband characteristics. The converter section is the most critical part of the whole transceiver, and is prewired and aligned by the factory for maximum performance. Sensitivity is 0.5 $\mu$V for 10 dB signal-plus-noise to noise ratio.

The output from the converter chassis is fed into the second mixer through a coaxial cable. The triode section of a 6EA8 is used as a tunable oscillator to provide injection for the second mixer; the second mixer is the pentode section of the same 6EA8. The oscillator circuit is a temperature-compensated shunt-fed Hartley circuit that tunes from 32 to 36 MHz; output from the second mixer is at 1650 kHz.

The 1650-kHz i-f strip in the TR-108 transceiver uses two high-gain 6BZ6 amplifiers and three double-tuned i-f transformers. The double-tuned i-f transformers provide a simple way to obtain narrow bandwidth. The selectivity is 6 dB down at the 8-kHz points. Image rejection is rated at 55 dB and i-f rejection is better than 50 dB.

The dual-diode 6AL5 used in the de-

tector/automatic noise limiter stage is very sensitive to low-level signals. The first diode section of this tube functions as a standard a-m detector. The agc voltage is derived from this section, filtered and fed back to the control grid of the i-f amplifier stages.

The second-diode section of the 6AL5 is used as a series-gate noise limiter. This-type of noise limiter introduces negligible distortion, even under strong-signal conditions with high-level modulation—most limiters distort the desired signal when limiting begins. The series-gate circuit is self-adjusting over a wide range of signal levels so the limiting level is compatible with the amplitude of the incoming signal.

One triode section of a 12AX7 is used for the audio amplifier to drive the 6L6 audio power stage; the other triode section of the 12AX7 serves as a microphone preamp. The audio output transformer has one primary and two secondaries; the primary is matched to the 6L6 audio power amplifier. In the receiver mode, one of the secondary windings is used to drive the speaker. A front panel jack allows headphone operation or an external speaker. The second audio-transformer secondary winding is used for modulation. The frequency response of the audio stage is engineered to cover from 300 to 3000 Hz, the speech range required for maximum intelligibility.

## transmitter

The transmitter may be crystal controlled with crystals in the 8-MHz range, or an external vfo may be used. The Knight-kit V-107 vfo is designed to go along with the TR-108 and its six-meter cousin, the TR-106. The 6CL6 oscillator/tripler stage is operated as a Colpitts oscillator with the plate circuit tuned to the third harmonic of the grid circuit. For vfo operation, the tube serves as an amplifier/buffer and tripler. Excellent vfo isolation is provided by the pentode.

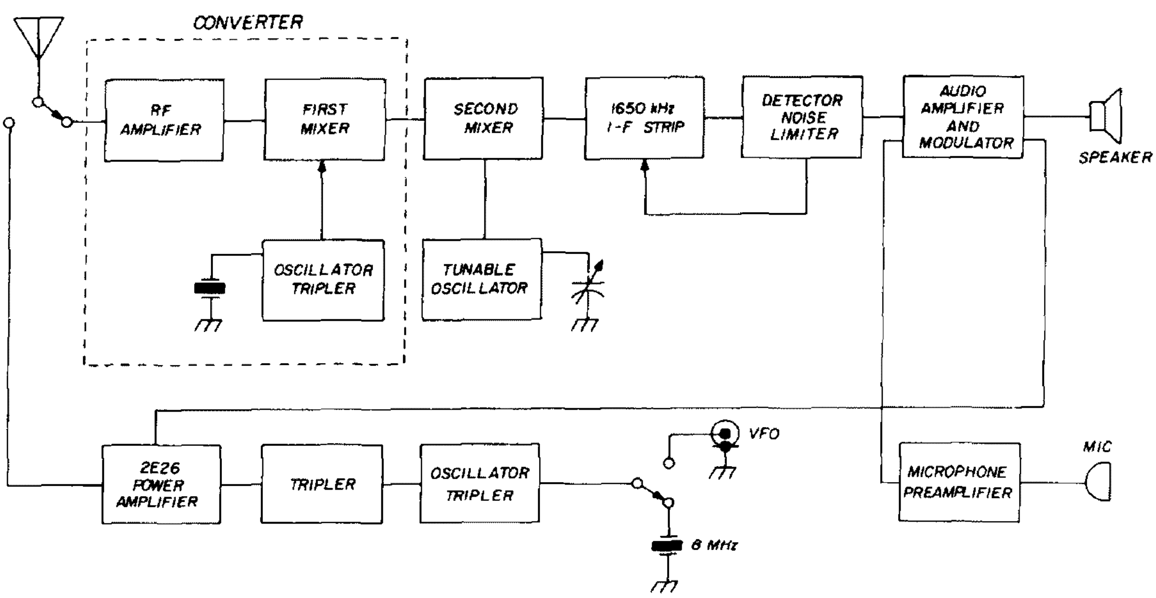



**fig. 1. Block diagram of the TR-108 two-meter transceiver.**

The output from the 6CL6 drives a 6BQ5 tripler stage. This small beam-power tube drives the 2E26 power amplifier to its full capacity. The compact 2E26 stage runs at 15 watts input. The output circuit is a conventional pi network designed to work into 30 to 90 ohms with a vswr less than 3:1.

The Knight-kit engineers gave harmonic generation very careful consideration when they designed the TR-108 in an effort to eliminate problems with harmonic radiation. A double-tuned transformer is used to couple energy from the 6BQ5 tripler to the final; this reduces harmonics and provides adequate drive over the complete two-meter band. The

output pi network furnishes further harmonic suppression and the wrap-around case shields the entire unit so undesired radiation is minimized.

## power supply

The power supply is one of the most interesting circuits in the transceiver since it is designed to operate from either 117 Vac or 12 Vdc. Switching from one power source to the other is automatically accomplished by using the proper power plug.

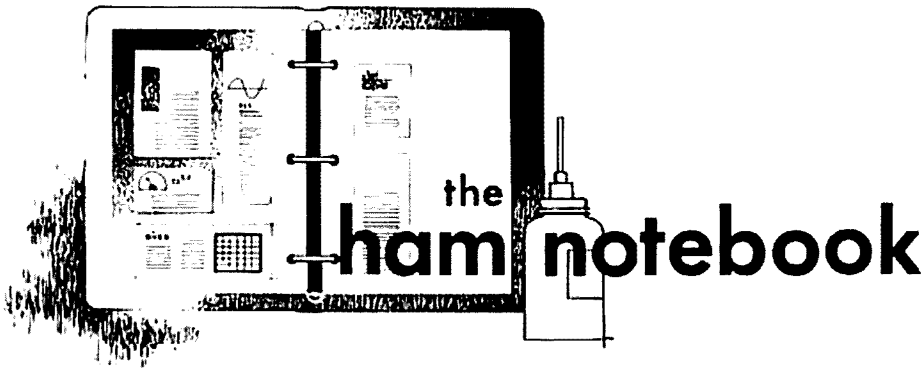
In the dc-to-dc-converter mode, the supply uses a two-transistor oscillator to switch the primary 12-volt dc power on and off at about 90 Hz: the resultant ac is stepped up by the transformer. In the ac mode, the transformer functions as a normal step-up type. The high voltage dc is provided by a voltage-doubler circuit across the transformer secondary. Power for an external vfo is available on the back panel.

## summary

The transceiver is fairly easy to assemble and with the excellent instructions furnished with the kit, only a few evenings work is required to do the job. Plenty of lineup and tuneup instructions are provided as well as hints for operating on the vhf bands. The instruction manual also includes information on building a simple antenna, mobile installation and mobile noise suppression, as well as some excellent dope on television interference and how to cure it.

Performance is right up to the manufacturer's rating. You're not going to work any moonbounce or meteor scatter with 15 watts input, but with a good antenna and patience, you may be able to work a little DX when the band opens up. In the meantime, the little TR-108 is great for local ragchews, DX nets and a-m repeaters. The receiver is more than sensitive—the 15-watt input rating of the transmitter is the limiting factor as far as maximum range is concerned. You can be sure you'll hear the other guy if he can hear you.

ham radio




## repairing broken coax connectors

The insulation in SO-239 coax chassis connectors sometimes breaks and falls out after a period of use. While a new connector costs only 59c, it is often far easier and quicker to repair a broken one than replace it. Set the piece of gear on the bench so the connector is upright. With a pair of long nose pliers squeeze the four parts of the inner connector sleeve together and fill the inner part with melted candle wax. Epoxy resin is used to replace the original broken and lost insulating material. After the epoxy hardens overnight remove the candlewax with a small screwdriver or with a 5/32" twist drill, and the SO-239 will be permanently repaired.

Robert B. Kuehn, WOHKF

## antenna dimensions

Here's a chart of antenna dimensions that should be handy if you're getting ready to put up a new sky wire. This is part of a computer listing I ran that covered the spectrum from 3.5 to 30 MHz with readouts for each 50 kHz change in frequency.* The first column of the chart, **table 1**, is the operating frequency in MHz. The following six columns give the dimensions of 5/8 wavelength, 1/2 wavelength, 1/4 wavelength, the length of one side of a cubical quad, the length of an inverted vee, and the proper radial length for a ground-plane antenna, all lengths given in feet. To convert from hundreths of feet to feet-and-inches use the chart in **table 2.**

Jim Barcz, WA9JMY

*A copy of the complete computer printout, from 3.5 to 30 MHz in 50 kHz steps, with dimensions to six decimal places, is avaiable from *ham radio* for $1.00.

**table 2. Conversion chart from hundredths of feet to inches. From table 1 an inverted-vee antenna on 3.9 MHz is 118.974 feet long. This is equivalent to 118 feet, 11.64 inches (after rounding off to 118.97 feet).**

| | 0 | 1 | 2 | 3 | 4 | 5 | 6 | 7 | 8 | 9 |
|---|---|---|---|---|---|---|---|---|---|---|
| 0 | 0.00 | 0.12 | 0.24 | 0.36 | 0.48 | 0.60 | 0.72 | 0.84 | 0.96 | 1.08 |
| 0.1 | 1.20 | 1.32 | 1.44 | 1.56 | 1.68 | 1.80 | 1.92 | 2.04 | 2.16 | 2.28 |
| 0.2 | 2.40 | 2.52 | 2.64 | 2.76 | 2.88 | 3.00 | 3.12 | 3.24 | 3.36 | 3.48 |
| 0.3 | 3.60 | 3.72 | 3.84 | 3.96 | 4.08 | 4.00 | 4.32 | 4.44 | 4.56 | 4.68 |
| 0.4 | 4.80 | 4.92 | 5.04 | 5.16 | 5.28 | 5.40 | 5.55 | 5.64 | 5.76 | 5.88 |
| 0.5 | 6.00 | 6.12 | 6.24 | 6.36 | 6.48 | 6.60 | 6.72 | 6.84 | 6.96 | 7.08 |
| 0.6 | 7.20 | 7.32 | 7.44 | 7.56 | 7.68 | 7.80 | 7.92 | 8.04 | 8.16 | 8.28 |
| 0.7 | 8.40 | 8.52 | 8.64 | 8.76 | 8.88 | 9.00 | 9.12 | 9.24 | 9.36 | 9.48 |
| 0.8 | 9.60 | 9.72 | 9.84 | 9.96 | 10.08 | 10.20 | 10.32 | 10.44 | 10.56 | 10.68 |
| 0.9 | 10.80 | 10.92 | 11.04 | 11.16 | 11.28 | 11.40 | 11.52 | 11.64 | 11.76 | 11.88 |
| 1.0 | 12.00 | 12.12 | 12.24 | 12.36 | 12.48 | 12.60 | 12.72 | 12.84 | 12.96 | 13.08 |

table 1. Antenna dimensions; length in feet

| frequency (MHz) | length (feet) 5/8 λ | 1/2 λ | 1/4 λ | quad side | inverted vee | radials |
|---|---|---|---|---|---|---|
| 3.50 | 167.143 | 133.714 | 66.857 | 71.714 | 132.571 | 68.571 |
| 3.55 | 164.789 | 131.831 | 65.915 | 70.704 | 130.704 | 67.606 |
| 3.60 | 162.500 | 130.000 | 65.000 | 69.722 | 128.889 | 66.667 |
| 3.65 | 160.274 | 128.219 | 64.110 | 68.767 | 127.123 | 65.753 |
| 3.70 | 158.108 | 126.486 | 63.243 | 67.638 | 125.405 | 64.865 |
| 3.75 | 156.000 | 124.800 | 62.400 | 66.933 | 123.733 | 64.000 |
| 3.80 | 153.947 | 123.158 | 61.579 | 66.053 | 122.105 | 63.158 |
| 3.85 | 151.948 | 121.558 | 60.779 | 65.195 | 120.519 | 62.338 |
| 3.90 | 150.000 | 120.000 | 60.000 | 64.359 | 118.974 | 61.538 |
| 3.95 | 148.101 | 118.481 | 59.241 | 63.544 | 117.468 | 60.759 |
| 4.00 | 146.250 | 117.000 | 58.500 | 62.750 | 116.000 | 60.000 |
| 7.00 | 83.571 | 66.857 | 33.429 | 35.857 | 66.286 | 34.286 |
| 7.05 | 82.979 | 66.383 | 33.191 | 35.603 | 65.816 | 34.043 |
| 7.10 | 82.394 | 65.915 | 32.958 | 35.352 | 65.352 | 33.803 |
| 7.15 | 81.818 | 65.455 | 32.727 | 35.105 | 64.895 | 33.566 |
| 7.20 | 81.250 | 65.000 | 32.500 | 34.861 | 64.444 | 33.333 |
| 7.25 | 80.690 | 64.552 | 32.276 | 34.621 | 64.000 | 33.103 |
| 7.30 | 80.137 | 64.110 | 32.055 | 34.384 | 63.562 | 32.877 |
| 14.00 | 41.786 | 33.429 | 16.714 | 17.929 | 33.143 | 17.143 |
| 14.05 | 41.637 | 33.310 | 16.655 | 17.865 | 33.025 | 17.082 |
| 14.10 | 41.489 | 33.191 | 16.596 | 17.801 | 32.908 | 17.021 |
| 14.15 | 41.343 | 33.074 | 16.537 | 17.735 | 32.792 | 16.961 |
| 14.20 | 41.197 | 32.957 | 16.479 | 17.676 | 32.676 | 16.901 |
| 14.25 | 41.053 | 32.842 | 16.421 | 17.614 | 32.561 | 16.842 |
| 14.30 | 40.909 | 32.727 | 16.364 | 17.552 | 32.448 | 16.783 |
| 14.35 | 40.767 | 32.613 | 16.307 | 17.491 | 32.334 | 16.725 |
| 21.00 | 27.857 | 22.286 | 11.143 | 11.952 | 22.095 | 11.429 |
| 21.05 | 27.791 | 22.233 | 11.116 | 11.924 | 22.043 | 11.401 |
| 21.10 | 27.725 | 22.180 | 11.090 | 11.896 | 21.991 | 11.374 |
| 21.15 | 27.660 | 22.128 | 11.064 | 11.868 | 21.939 | 11.348 |
| 21.20 | 27.594 | 22.075 | 11.038 | 11.840 | 21.887 | 11.321 |
| 21.25 | 27.529 | 22.024 | 11.012 | 11.812 | 21.835 | 11.294 |
| 21.30 | 27.465 | 21.972 | 10.986 | 11.784 | 21.784 | 11.268 |
| 21.35 | 27.400 | 21.920 | 10.960 | 11.756 | 21.733 | 11.241 |
| 21.40 | 27.336 | 21.869 | 10.934 | 11.729 | 21.682 | 11.215 |
| 21.45 | 27.273 | 21.818 | 10.909 | 11.702 | 21.632 | 11.189 |
| 28.00 | 20.893 | 16.714 | 8.357 | 8.964 | 16.571 | 8.571 |
| 28.10 | 20.819 | 16.655 | 8.327 | 8.932 | 16.512 | 8.541 |
| 28.20 | 20.745 | 16.596 | 8.298 | 8.901 | 16.454 | 8.511 |
| 28.30 | 20.671 | 16.537 | 8.269 | 8.869 | 16.396 | 8.481 |
| 28.40 | 20.599 | 16.479 | 8.239 | 8.838 | 16.338 | 8.451 |
| 28.50 | 20.526 | 16.421 | 8.211 | 8.807 | 16.281 | 8.421 |
| 28.60 | 20.444 | 16.364 | 8.182 | 8.776 | 16.224 | 8.392 |
| 28.70 | 20.383 | 16.306 | 8.153 | 8.746 | 16.167 | 8.362 |
| 28.80 | 20.313 | 16.250 | 8.125 | 8.715 | 16.111 | 8.333 |
| 28.90 | 20.242 | 16.194 | 8.097 | 8.685 | 16.055 | 8.304 |
| 29.00 | 20.172 | 16.138 | 8.069 | 8.655 | 16.000 | 8.276 |
| 29.10 | 20.103 | 16.082 | 8.041 | 8.625 | 15.945 | 8.247 |
| 29.20 | 20.034 | 16.027 | 8.014 | 8.596 | 15.890 | 8.219 |
| 29.30 | 19.966 | 15.973 | 7.986 | 8.567 | 15.836 | 8.191 |
| 29.40 | 19.898 | 15.918 | 7.959 | 8.537 | 15.782 | 8.163 |
| 29.50 | 19.830 | 15.864 | 7.932 | 8.508 | 15.729 | 8.136 |
| 29.60 | 19.764 | 15.811 | 7.905 | 8.480 | 15.676 | 8.108 |
| 29.70 | 19.697 | 15.758 | 7.879 | 8.451 | 15.623 | 8.081 |

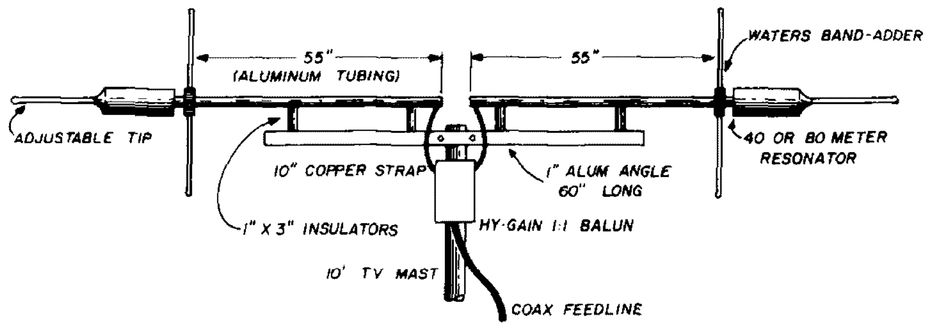



**fig. 1. Multiband dipole designed for wooden boats by W2INS. This antenna would also be suitable for portable and apartment stations.**

## portable all-band antenna

Although the antenna shown in **fig. 1** was designed primarily for operation on large wood boats, it looks ideal for any amateur with limited space. Vertical antennas are virtually unmanageable on large non-metal boats, and must be constantly retuned, matched and re-resonated; this all-band miniature dipole is relatively non-critical, easy to tune and easily matched. Although it can be rotated to take advantage of any inherent directivity, most users don't go to the trouble. The antenna is tuned up on the lowest band of operation (either 40 or 80 meters) with a grid dipper and antennascope. Operation on the higher bands is provided by the two Waters Mobile Band-Adders. The original design of this antenna is attributed to W2INS, and it has been built by a number of boat enthusiasts with excellent success.

## elmac chirp and drift

Even though the Elmac AF-67 is fifteen years old, it still has many fine attributes. In fact, I just bought one. It worked fine on a-m, and it's tailored speech characteristics are as desirable today as when *it originally* was designed. However, in its original form it had two serious flaws that limited it's usefulness: a chirpy cw signal and unacceptable drift on 15 and 20 meters.

The chirp was cured in my unit by regulating the supply to the buffer/multipler stages. A 420-ohm dropping resistor and a 10k bleeder along with an OB2 regulator in series with an OA3 provide the proper operating voltages (see **fig. 2**). Keying with the new regulated supply is clean and chirpless.

Frequency drift was noticed only on 15 and 20 meters. A 35-pF variable capacitor, C15, is unique on these bands,

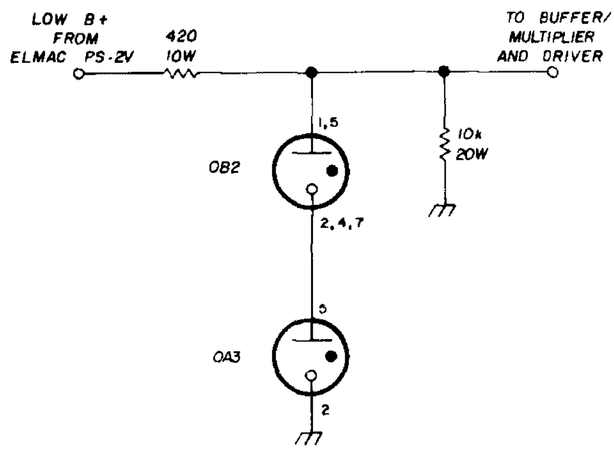
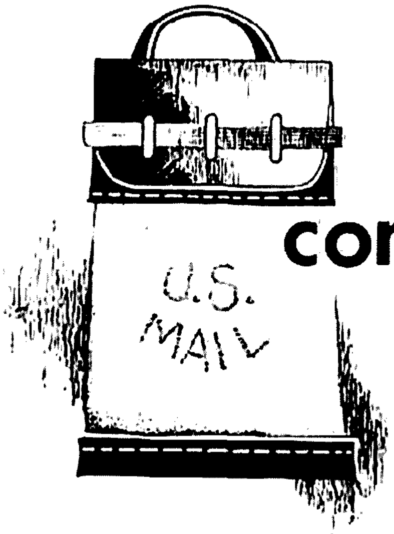



**fig. 2. Regulated supply eliminates chirpy cw signal from Elmac AF-67.**

so it was suspect. Sure enough, when it was replaced with an ordinary 35-pF trimmer the drift disappeared. I mounted the trimmer on the bottom plate of the vfo module because the original position was almost inaccessible.

I carefully examined the faulty capacitor and found that only one leaf of the rotor shaft contact spring was soldered in place—the other leaf maintained electrical contact through pressure. I suspect that some type of diode action was occuring between the pressure spring and the mounting nut, but because of the inaccessibility of the original mounting position, I didn't put the unit back in the transmitter to run further tests.

**George Hirshfield, W5OZF**




# comments

## multiband antennas

Dear HR:

First, allow me to congratulate you for a fine magazine. Through it many valuable and useful data become available to those of us who otherwise have no access to it; e.g. specifically practical design data for semiconductor high-frequency applications. If you're not employed in the business, as I'm not, such information is hard to come by. Although a poor schoolteacher can hardly afford to do much with it yet, I do find it invaluable in maintaining a modern perspective in electronics.

However, I do have reservations about some things I read in your magazine. Take for instance Mr. Orr's, "Simple Dual-band Antennas" in the March issue. Now, far be it to deny Mr. Orr's technical veracity; I have long found his writings both stimulating and correct. But, on the basis of some grassroots antenna experience I'd like to challenge the practical radiating efficiency of his half-wave radiator, folded-into a quarterwave space. I don't think that he or anyone else can do it.

Living as I do in one of those relatively small-lot suburban areas this business of getting up a good antenna for the 80-meter cw band is a live, pragmatic issue. In facing this issue at my own place, and that of my friends, I have found that a doublet-type antenna with a total *radiating* length of about four-tenths of a wavelength (100-foot wire on 80 meters, or about fifty feet on 40 meters) is the very minimum length that will produce satisfactory practical communication results. This, of course, assumes a means of resonating the wire to the operating frequency. Furthermore, this is independent of the feeding technique providing it is properly done.

Now, it's a plain fact that *any* unshielded conductor carrying appreciable RF current will radiate. James Clerk Maxwell told us as much back in 1870 or so. But to merely radiate, and to radiate *effectively* under typical amateur conditions in a ham's backyard, represent two different things.

Let's not talk microvolts per meter or things like that—I'll define "effective radiation" from an antenna as occurring when a typical fifty-watt (dc input) amateur station in the cw portion of the band can successfully compete with normal QRM conditions, achieving at least fifty percent "comebacks" on calls properly made under any but contest conditions. Unless Mr. Orr's location is considerably better than mine I'll respectfully bet him a drink that his antenna described in fig. 2, page 19 (March, 1970 issue) will *not* meet this test consistently unless his antenna is more than thirty feet above "dirt." (Most of my friends and myself have to work with antennas considerably lower than this.)

Mr. Orr makes much of the satisfactory swr he achieves with his antenna. This is nice, I suppose, but he could achieve an swr of unity very easily merely by hanging a 50-ohm, non-inductive resistor across the far-end of his RG-8/U. So what? Swr near unity is a desideratum, no doubt, but it is no guarantee of a good signal as any experienced radioman knows. What counts is, "stirring up the ether" as we used to say back in the pre-relativity days.

If Mr. Orr would take the wires he "folds back," and drop them vertically down to within five feet of the ground at each end (to make the total *wire* length at least 100 feet), then couple a *good* open-wire line to the center instead

of that pretty but lossy coax I'll bet he would stir-up the ether much more. Of course, he'll need a good antenna tuner at the bottom end of his feedline but that's no problem for a good man. He can then run a short piece of coax to his antenna relay. Then he'll be in business for blood instead of kicks. Furthermore, if he makes the tuner coils plug in he can tune his antenna for other higher frequency bands too with remarkably good results. (Why are so many "modern" radiomen apparently allergic to the good old tuned, open-wire line? What's so sacred about coax?)

Let me put results where my mouth is: last winter I was interested in some QRP work with a small 8-watt (input) transmitter on 80-meter cw. My first attempts were with an approximately quarter-wave-long doublet. I made a contact or two but not much fun. Then I hung about fifteen feet of wire on each end of the doublet. One would think that I had jacked up the eight-watter and put a hundred watter in its place! I soon worked all districts with my pipsqueak and had *many* long, solid ragchews. Lots of fellows thought I was kidding them when I told them my input was only eight watts. So, who is correct here, James Clerk Maxwell, Mr. Orr, or yours truly? Probably all of us somehow or other. But I hope I've stirred-up a little thought—and maybe an anti-coax revolution.

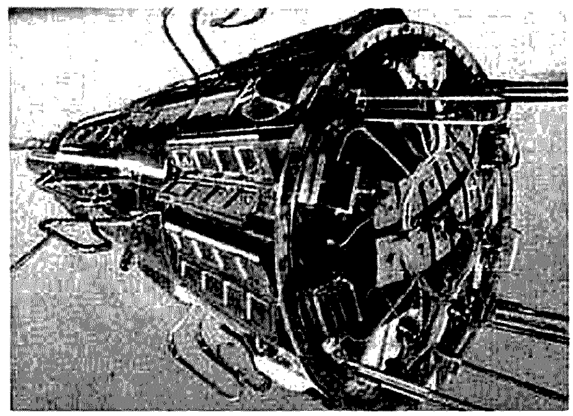
**C. R. Rockey, W9SCH/W9EDC**
**Deerfield, Illinois**

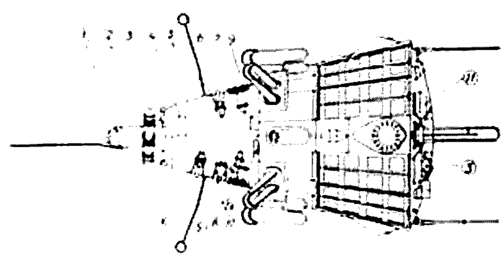
## make mine canadian club

*Of course, Mr. Rockey is correct. And so are James Clerk Maxwell and I. All practical antennas are a series of compromises. The finesse comes into play in selecting compromises available to best fit the situation.*

*The antenna in question, a compact folded dipole, works. As my first witness, enclosed is a photo and drawing of the Russian Sputnik III space satellite which bristles with folded monopoles (eight in all). A similar form of antenna is often used for glide-slope detection on jet aircraft (see* Antenna Engineering Handbook, *first edition by Jasik, McGraw Hill, N. Y., section 27-5).*

*Personally, I've used an 80-meter sixty-five foot folded dipole about 20 feet in the air for local skeds (up to 500 miles or so using 180 watts PEP input). No difficulty was experienced in laying down a strong signal with the abbreviated antenna.*





*Some years ago I used a vertical monopole version of this antenna squished down to 22 feet with four full-size radials. Using a kilowatt on 80 meters WAC was achieved including many European Contacts—which really separate the men from the boys on the West Coast (antennawise).*

*Finally, I take mild issue to Mr. Rockey's claim that a 100 foot antenna is the very minimum length that will produce satisfactory results on 80 meters. Look about at the eight-foot antennas that are used every day: they are called* mobile whip antennas. *(You can use two of them back to back for a pretty good sixteen foot dipole, too!.)*

*I demur from entering the coax-versus open wire line discussion. Using open*

*wire line, in my opinion, is like washing your feet with your socks on.*

*On second thought, I'll reserve that drink offer until I meet Rock. He sounds like a great guy. Maybe we'll get together some day for a chat and I'll order two CC on the rocks. Bottoms up!.*

**William I. Orr, W6SAI**

## a-m modulator

**Dear HR:**

In reference to Mr. Hall's article "A Different Approach to Amplitude Modulation," (*ham radio,* February, 1970) it should be pointed out that some undesirable aspects are associated with this type of approach. Although this modulation technique does eliminate the modulation transformer and the conventional class-B modulator transistors, it is interesting to examine the trade-offs involved.

Basically, the method described consists of modulating the power supply series regulator transistor. The problem arises from the unusually high dissipation this system requires of the regulator transistor. Unlike conventional modulation where the rf amplifier collector current flows through the low resistance secondary of the modulation transformer, this method requires that the entire rf amplifier current flow through a regulator transistor which must have a voltage drop across it greater than the supply voltage to the rf amplifier. This means that the average regulator transistor dissipation will always exceed the entire input to the transmitter rf amplifier.

Although the article makes reference to the fact that efficiency is low, the figures quoted are somewhat misleading because they are bases on 100% sine-wave modulation. In practical use involving voice modulation and normal pauses between words, the quiescent dissipation becomes more significant from a heat producing standpoint. Since the idling current for conventional class-B modulator transistors is quite low, Mr. Hall's method would require nearly twice the input power in the quiescent state. This is a situation similar to running the modulator transistors of a conventional system in class A.

Although the regulated power supply would probably be used with either method, with conventional a-m the modulator would not normally be supplied from the regulated output. Therefore, the regulator transistor dissipation would be modest since the power transformer voltage would be selected such that only a few volts are dropped across the regulator transistor.

The fact that this amplitude modulation method requires twice the normal supply voltage and, therefore, consumes twice the amount of power (idling) would probably preclude its use with battery operated equipment. The choice of this circuit for ac operated (or mobile) equipment should be quided by whether it is practical to dissipate the additional heat produced. The space savings facilitated by eliminating the modulation transformer and transistors may conceivably be offset by the additional heat sink area required to satisfy regulator transistor temperature requirements.

**Jerry Manikowski, W9JGV**
**Hanover Park, Illinois**

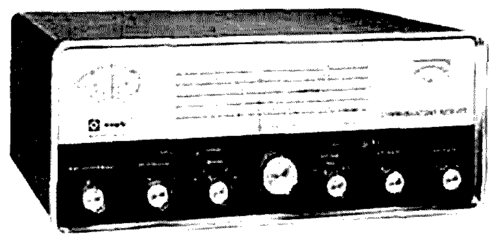
*In reply to Mr. Manikowski's remarks about my article, I wish to thank him for bringing up the fact that the efficiencies mentioned in the article are misleading because they are based on 100% sine wave modulation. His point is well taken, and I am guilty of an oversight there.*

*The "usually high dissipation" required of the regulator transistor is the price that must be paid for elimination of the conventional modulator. It was not my intention to imply that this modulation approach offered something for nothing; rather, it is presented for its novelty and because it requires no modulation transformer. It is hoped that some builders will find the scheme useful in some applications, as indeed one has.*

**Courtney Hall, WA5SNZ**
**Dallas, Texas**

# new products

## general-coverage receiver

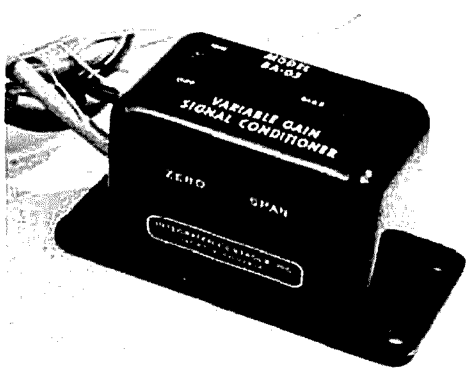


The new Knight-Kit model R-195 solid-state communications receiver kit tunes the international and domestic shortwave bands, marine weather and navigational beacon bands plus the standard a-m broadcast band. (Frequencies are 200-420 kHz, 550-1800 kHz, 1.8-12 MHz.) The receiver includes a tuned rf stage for high sensitivity and low noise, and the large illuminated slide-rule tuning dial makes tuning easy.

Field-effect transistors are used in the front end for high sensitivity with low cross modulation, and ceramic i-f filters provide selectivity and interference rejection. Sensitivity is better than 2 µV for 10 dB signal-plus-noise to noise. Selectivity is 4.5 kHz at the 6 dB points. Other features include series-gate automatic noise limiter, avc, built-inbfo and product detector, remote receiver muting connections, built-in speaker and headphone jack.

The R-195 receiver kit is priced at $89.95. Knight-Kit's new modular concept makes assembly of this kit easy even if you've never built a kit before. Most parts are already soldered to the printed-circuit boards, and all critical adjustments have been made at the factory. All the builder has to do is follow the detailed step-by-step instructions and solder the connections between circuit boards. Full description of this new receiver can be found in Allied 1970 catalog, sent free on request from Allied Radio Corporation, 100 N. Western Avenue, Chicago, Illinois 60680.

## instrumentation amplifier

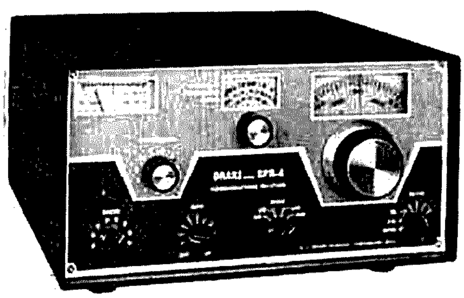



A low-cost self-powered IC signal conditioner with continuously variable voltage gain from one to greater than 1000 has been announced by Integrated Controls, Inc. Input impedance is greater than 100 megohms on the lower gain ranges, and the input overvoltage capability is ±30 Vdc continuously. The output is short-circuit proof, and current limits at 20 mA. The gain-bandwidth product is 700 kHz, low-frequency noise is typically 10 µV, and gain stability is typically better than 100 ppm/°C. Operating temperature range is zero to +70° C. The mercury battery pack provided permits greater than 1000 hours of continuous operation exclusive of transducer excitation. $85 from Integrated Controls, Inc., P. O. Box 17296, San Diego, California 92117.

## test equipment catalog

A new 44-page test equipment catalog from Tucker Electronics describes over 2000 different test instruments and microwave components for sale or rent. Most of the listed items are either new-surplus or used-reconditioned; all are calibrated to the manufacturer's original specifications by a calibration laboratory (traceable to N.B.S.). Included in the catalog are voltmeters, transfer standards, signal generators, oscilloscopes plus many others; manufacturers include Dumont, Fluke, General-Radio, Hewlett-Packard, Polarad, Tektronix and others. For a copy of catalog no. 18, write to Tucker Electronics Company, P. O. Box 1050, Garland, Texas 75040.

## communication receiver



The new Drake SPR-4 communications receiver is an all solid-state unit that may be programmed to suit your present and future interests. The front-end of this new receiver uses a dual-gate mosfet that provides signal-handling characteristics that are said to be superior to the best tube receivers. The all solid-state design, of course, has the advantages of low power consumption, mechanical and thermal stability and reliability.

The SPR-4 can be programmed with accessory crystals for 23 ranges (each tuning a 500-kHz band) from 0.5 to 30 MHz plus 150 to 500 kHz. Crystals supplied with the receiver allow coverage on these ranges: 150-500 kHz, 0.5-1.0 MHz, 1.0-1.5 MHz, 6.0-6.5 MHz, 7.0-7.5 MHz, 9.5-10 MHz, 11.5-12 MHz, 15-15.5 MHz, 17.5-18 MHz and 21.5-22 MHz. These ranges cover the most popular short-wave listening frequencies and include aircraft radio and weather, marine ship and shore stations, high-frequency communications, WWV standard time signals, foreign broadcast, CB and amateur radio.

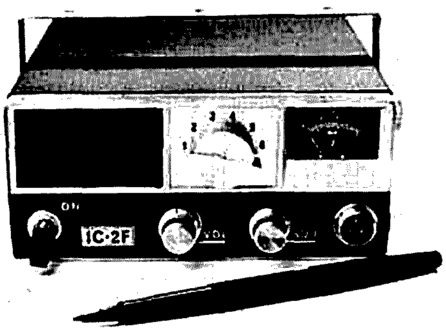
This new receiver features avc on a-m, cw and ssb with time constants selected for optimum effectiveness on each mode. Audio output is held constant within 3 dB over a 100-dB range of input signals. Sensitivity on ssb and cw is 0.25 $\mu$V for 10 dB signal-plus-noise to noise ratio. Hum and noise are more than 60 dB below rated output. Accessories include a matching speaker, crystal calibrator, noise blanker, loop antenna, and transceive adapter that allows full transceive operation with Drake T-4B and T-4XB transmitters. The model SPR-4 is priced at $379.00. For more information, or the name of your local dealer, write to R. L. Drake Company, 540 Richard Street, Miamisburg, Ohio 45342

## eimac application bulletin

The new Eimac 4CX600 family of ceramic-metal tetrodes is designed to meet the demands of modern communications systems. These new tubes are ruggedized, compact, radial-beam tetrodes for use up to 890 MHz. Closely controlled parameters plus state-of-the-art assembly and testing permit an extremely high-gain bandwidth product and low intermodulation distortion level to be achieved simultaneously.

Eimac's Application Bulletin number 14 covers these tubes in detail and includes circuits for a tunable 140-250-MHz stripline amplifier, a 432-MHz cavity amplifier, a 150-MHz class-B amplifier, as well as many others. This bulletin, which normally sells for $1.50, will be available free of charge to the first 100 readers of *ham radio* who write to Bill Orr, W6SAI, Advertising Manager, Eimac Division of Varian, 301 Industrial Way, San Carlos, California 94070

## vhf fm transceiver




The new Varitronics IC-2F fm transceiver for two meters (and IC-6F for six meters) provides a full 20 watts input with 12 or more watts out. New innovations in these units include APC (automatic protection circuit) which consists of an swr bridge and two dc amplifiers that bias the power-amplifier and driver transistors off if the unit sees high vswr in the transmit mode. This saves the transistors from possible damage. In addition, direct access is provided to the discriminator so an outboard zero meter can be used, and all switching is done electronically, eliminating problems with relays. These new transceivers also include a dc input filter that does away with alternator hash, a squelch circuit that uses thermistors to eliminate temperature-induced squelch drift, a hinged chassis for easy servicing, test points brought up from boards for easy voltage, current and frequency checks, a low-pass output filter, and zener-regulated supply voltages for all oscillators.

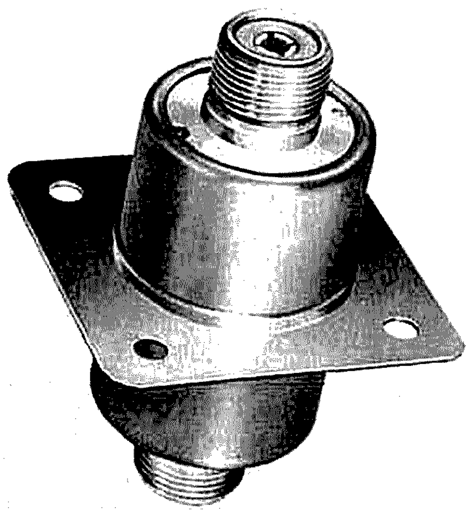
The receiver features a quiet fet front end with ceramic filters and integrated-circuit i-f stages. Actual sensitivity is below 0.2 $\mu$V for 20 dB quieting.

Units are supplied complete with crystals, microphone, mounting bracket, plugs and dc cabling. The only accessory is the ac power supply (IC-3P). This power supply incorporates a sensitive zero-type discriminator meter, and is designed to provide a stand for the transceiver. Hardware is provided to mate the transceiver and the ac power supply into one neat compact unit, if desired. For more information on the new IC-2F and IC-6F, write to Varitronics Incorporated, 3835 North 32nd Street, Suite 6, Phoenix, Arizona 85018.

## decoder/driver ic

Motorola's new BCD-to-decimal decoder/driver, the MC9860/9760, converts a four-bit complementary BCD code into a one-of-ten output with sufficient voltage to drive a neon-filled display tube. The high-voltage output transistors in the ICs allow direct operation of the display tube without background glow. With the MC9860/9760 decoder/driver, MC867/767 quad latch and MC880/780 decade up counter, designer can build a completely integrated readout system that is smaller, faster, and less expensive than conventional systems. Supplied in the 16-lead dual in-line plastic package, the MC976OP sells for $5.55 in small quantities. For more information, write to Technical Information Center, Motorola Semi-conductor Products, Inc,. Box 20924, Phoenix, Arizona 85036.

## lightning arrestor



A lightning arrestor originally developed for protecting military communications equipment is now available to radio amateurs. This new arrestor, the Dale Electronics LA-8A4C, provides permanent protection from lightning damage to your radio equipment. This arrestor has

been proven so reliable that Dale Electronics will pay you $2000 if a properly-installed LA-8A4C fails to protect your equipment from lightning damage. The ultra-reliable LA-8A4C can withstand five direct strokes of lightning, eliminates heavy static buildup and installs easily in any 52- or 72-ohm coaxial feedline. $19.95 is the standard price for the arrestor–the added protection of the $2000 warranty policy costs you nothing. Dale Electronics, Inc., P.O. Box 609, Columbus, Nebraska 68601.

## fm transceiver

Galaxy Electronics has announced a new two-meter fm transceiver, the FM-210. This new solid-state unit features a high-performance fet-front-end transceiver and 3-channel crystal-controlled independent transmit/receive selected by front panel controls, highly effective squelch system and a speech compressor for optimum intelligibility under adverse operating conditions. The FM-210 is designed for operation from 12- to 14-Vdc supplies; an optional power booster provides higher power operation from either 12-14 Vdc or 117 Vac. Total power drain in the receive mode, with the audio squelched, is 60 mA; during transmit the current drain is 500 mA.

Receiver sensitivity is 1 uV for 12 dB quieting; 0.5 uV for 12 dB SINAD. On transmit the unit runs 5 watts input (3 watts out), and with the optional power booster, 10 watts input (6 watts out). The power booster is actually a dc-to-dc converter that provides 24 Vdc to the final power amplifier transistor. The transceiver also includes adjustable bandwidth, 2 to 20 kHz, and an adjustable clipping filter that provides up to 30 dB clipping. The FM-210 is normally configured for wideband operation; narrow band is available on special order.

For more information on this new unit, write to Galaxy Electronics, 10 South 34th Street, Council Bluffs, Iowa. The transceiver is priced at $299.95; the optional power booster is $39.95.

## short circuits

Conversion of the Hallicrafters SR-160 to an SR-500 (to gain greater power output), *cannot* be done as described on page 66 of the February, 1970 issue. The biggest problem lies in the filament circuit—the 8236 tubes have 6-volt heaters and will be immediately zapped if the SR-160's 12-volt filament circuit isn't changed.

### high-frequency converter

The crystal oscillator stage in the high-frequency mosfet converter shown on page 30 of the January, 1970 issue is incorrect. Each of the crystals should be returned to the base of the 2N4124 oscillator stage, *not* to ground. Also, the unlabeled capacitor across L5 in the mixer drain circuit should be 170 pF.

### varactor modulator

In the frequency-modulated crystal oscillator in **fig. 10**, page 18 of the September, 1969 issue, the MV-833 varactor is installed backwards. The diode must be reverse biased for correct operation.

### 432-MHz ssb converter

The B+ supply to the crystal oscillator in fig. **1**, page 50 of the January, 1970 issue should come directly from the positive terminal of the OC5 voltage-regulator tube. Also, the oscillator B+ lead should be bypassed with a uhf-type feedthrough capacitor, and the cathode bias resistor in the 6J4 grounded-grid amplifier stage should be 56 ohms.

### ic noise blanker

The 47k resistor in the emitter circuit of the Schmit trigger in fig. 3, page 54 of the May, 1969 issue should be 47 ohms. The circuit will not trigger reliably unless this resistor is less than 150 ohms.

ham radio

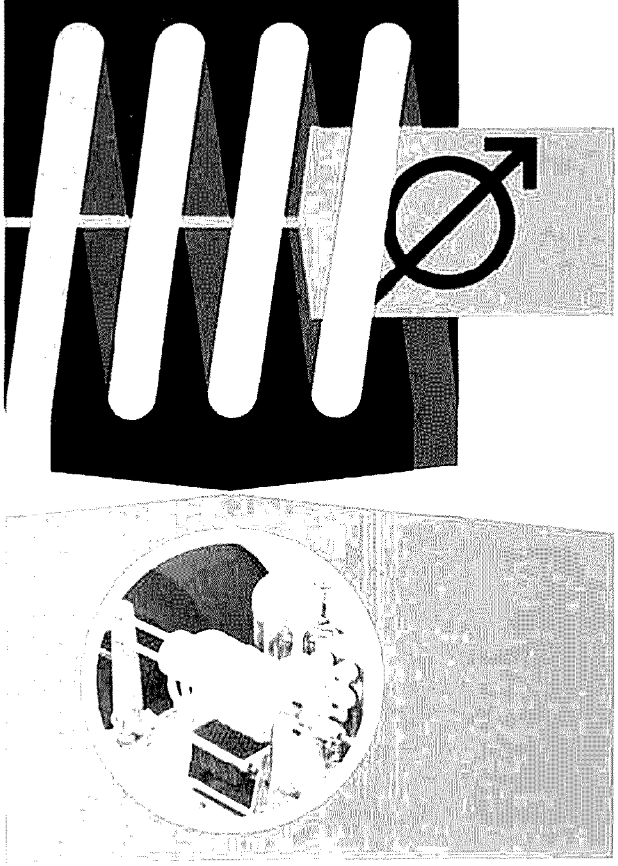
75 cents

focus on communications technology . . .

# ham radio magazine

JULY 1970

## INDUCTIVELY-TUNED HIGH-FREQUENCY TANK CIRCUIT



this month

july 1970
volume 3, number 7

**staff**

James R. Fisk, W1DTY
editor

Nicholas D. Skeer, K1PSR
vhf editor
J. Jay O'Brien, W6GDO
fm editor

A. Norman Into, Jr., W1CCZ
Alfred Wilson, W6NIF
James A. Harvey, WA6IAK
associate editors

Jean Frey
art director
Wayne T. Pierce, K3SUK
cover

T. H. Tenney, Jr. W1NLB
publisher

John B. Morgan, K1RA
advertising manager

**offices**

Greenville, New Hampshire 03048
Telephone: 603-878-1441

**ham radio** magazine is
published monthly by
Communications Technology Inc.
Greenville, New Hampshire 03048

Subscription rates, world wide:
one year, $6.00, three years, $12.00.
Second class postage
paid at Greenville, N. H. 03048
and at additional mailing offices

Foreign subscription agents:
United Kingdom:
Radio Society of Great Britain,
35 Doughty Street, London WC1, England.

All European countries:
Eskil Persson, SM5CJP, Frotunagrand 1,
19400 Upplands Vasby, Sweden.

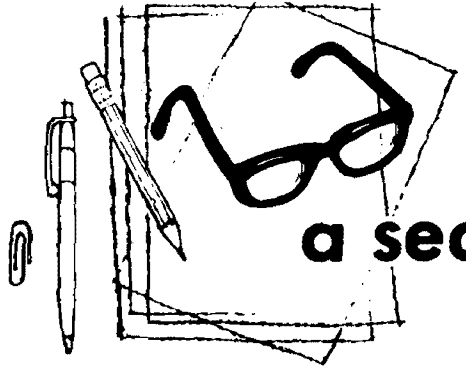

Printed by Wellesley Press, Inc.
Wellesley, Massachusetts 02181, U.S.A.

**ham radio** is available to the blind
and physically handicapped on magnetic tape
from Science for the Blind,
221 Rock Hill Road, Bala Cynwyd,
Pennsylvania 19440.
Microfilm copies of current
and back issues are available
from University Microfilms,
Ann Arbor, Michigan 48103.

Postmaster: Please send form 3579 to
**ham radio** magazine, Greenville,
New Hampshire
03048

**ham radio magazine**

# contents



# a second look

by jim fisk

**Do you remember the days** when the upper frequency limit of transistors was limited to a few hundred kHz? The days when the basement experimenter would sort through a box of new transistors hoping to find at least one device that he could use in a 455 kHz i-f strip? Some semiconductor scientists even doubted that high-frequency transistors could be manufactured. Then Philco came out with the SB-100 surface-barrier transistor and *practical* high-frequency rf circuits started appearing in the magazines. After that came Fairchild's Planar process, and the gradual but major changeover from germanium to silicon. Since then improved manufacturing techniques have resulted in transistors with higher frequency capability, lower noise output and larger power dissipation than we could have imagined–and the end is not in sight.

Microwave power transistors, although still too expensive for amateur applications, have been available for some time. However, for the most part power levels have been limited to the milliwatt range. Power limits have been slowly creeping up though, and this summer Fairchild is expected to market *10-watt* power transistors for operation at *2000 MHz.* It is reported that they will also introduce some new low-noise small-signal devices with very high gain-bandwidth products.

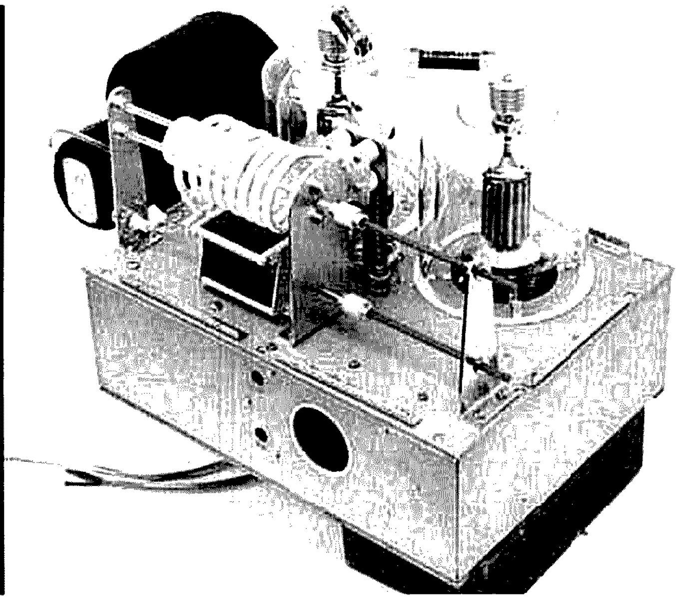
Although not too many hams venture into the world of the super-high frequencies, IBM's research division in Switzerland has come up with a gallium-arsenide field-effect transistor that exhibits more than 3-dB gain at 17,000 MHz. From data gained in their laboratories it appears that the maximum oscillation frequency for this device may be as high as 30,000 MHz. This is at least 2½ times higher than any current models.

The new gallium-arsenide transistor is called a mesfet–metal semiconductor field effect transistor, and uses a Schottky barrier gate, rather than the more familiar insulated-gate arrangement. A rectifying contact is established directly at the metal-semiconductor's surface. A chromium-nickel-gold sandwich, 40 millionths of an inch wide and 8 thousandths of an inch in diameter, forms a circular gate that completely surrounds the circular-shaped drain.

From another part of the semiconductor world comes a microwave diode packaged in a 1N23 rectifier cartridge that produces peak powers of 1.2 kW at 1100 MHz. A result of work at RCA, the device actually consists of five avalanche diodes in series. Avalanche diodes use drift time, that old bugaboo of vacuum tubes, to generate microwave power. Bunched current carriers drift through the solid-state crystal of the diode and deliver rf power by causing an external circuit current 180° out of phase with the voltage. These diodes are often referred to as IMPATT devices (from IMPact Avalanche Transit Time), and are like klystrons in some respects since both use transit time and current-carrier bunching in their operation.

Because of high cost, it's doubtful that any of these devices will see widespread amateur use in the immediate future, but the nature of semiconductor manufacturing is such that as techniques improve, yields go up, and prices come down. Remember when a new CK722 audio transistor wiped out a ten-dollar bill? Today you can buy an equivalent device for 39¢–and if you're willing to take a chance, $1 the bagful.

**Jim Fisk, W1DTY**
**editor**



# inductively-tuned high-frequency tank circuit

A method for achieving high efficiency in the "shadow region" between 14 and 54 MHz

William I. Orr, W6SAI, Eimac Division of Varian

Radio frequency amplifiers require a certain critical value of plate circuit impedance and Q for optimum performance at any frequency. Design deviations may lead to higher levels of intermodulation distortion or excessive harmonic radiation. While the design requirement may be quite tolerant in some cases, the mechanical assembly of the components and the choice of proper values become increasingly critical as the operating frequency nears the upper region of the hf spectrum. It is as though a "shadow region" exists that's too high for conventional lumped circuit components, yet is too low for conventional vhf linear and stripline techniques. The "shadow region" extends roughly from 27 through 54 MHz.

Above 27 MHz or so, the construction of a conventional high-power plate-tuning circuit having good Q and good efficiency can be vexing, as residual tube and circuit capacitances combine to assume a major portion of the tank circuit capacitance. It's possible, in fact, for this residual

capacitance to be much larger than specified for proper design considerations. The unusually high circuit capacitance may lead to unreasonable Q and high circulating tank current, resulting in poor over-all efficiency and excessive heat loss in the tank circuit.

## inductive tuning

To achieve good circuit efficiency and proper Q in the upper portion of the hf spectrum, it is convenient to resort to a different mechanical configuration than is commonly used at lower frequencies. One way to overcome problems of efficiency and Q is to reduce residual circuit capacitance to an absolute minimum by removing the cause of the largest portion of this unwanted capacitance: the tank tuning capacitor. Circuit resonance can then be established by including a fixed capacitance combined with a variable inductor.

The inductor can be a fixed, high-Q coil having a low-loss shorted turn introduced into one end. As the turn is moved within the coil, coil inductance is reduced, and resonance is established by correctly positioning the shorted turn (fig. 1). The rf current in the shorted turn is high compared to the coil current; however, if the turn is of homogeneous structure and low-resistance material, turn losses will be small.

Shown in the photographs is a commercial 5-kW input PEP linear amplifier designed for any 500-kHz range between 27 and 54 MHz. The plate tank is a conventional pi network, inductively tuned by a shorted turn within the plate coil. The turn (or "slug") is moved into and out of the coil by a lead screw driven from a counter dial mounted on the amplifier panel. Tank circuit values were derived from pi network charts.[1]

The amplifier uses a pair of parallel-connected 3-1000Z high-mu triodes in a grounded-grid, cathode-driven circuit with zener diode bias.[2] The combined output capacitance of the tubes is approximately 15 pF. Stray circuit capacitance from plate to ground is less than 10 pF, which provides a minimum input capacitance (C1) for the pi network of about 25pF.

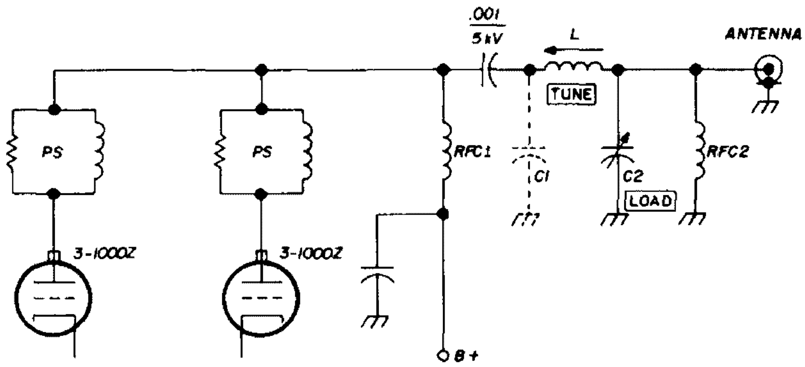



**fig. 1. Inductively tuned tank circuit using conventional component values. C1 is the residual circuit capacitance plus tube output capacitance. Resonance is achieved by a variable shorted turn moved inside L.**

Additional capacitance can be added for operation at lower frequencies. Ceramic or vacuum padding capacitors can be used for this purpose.

## construction

Tank-circuit inductance is calculated for the low-frequency end of the tuning range in the usual manner. As the shorted turn is driven into the plate inductor, its inductance decreases, and resonant frequency increases. A range of about 500 kHz or so can be obtained with this assembly.

Construction details are shown in **fig. 2.** The shorted turn is a section of seamless copper water pipe 1-¾ inches O.D. by 3 inches long. Discs of insulating material are cut to fit into the ends of the pipe and are held in position with small pins or rivets. A

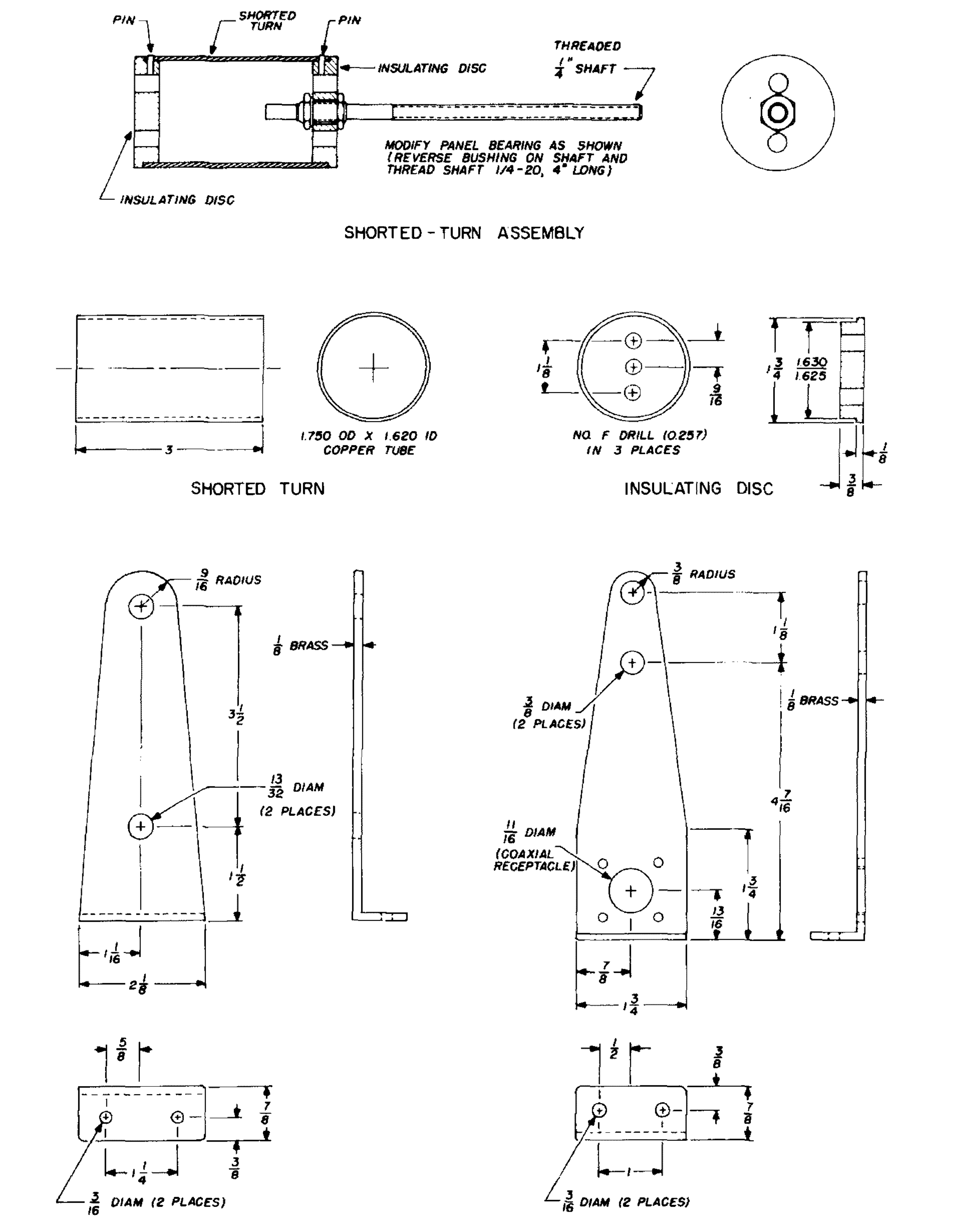



*fig. 2. Mechanical details of the shorted-turn assembly.*

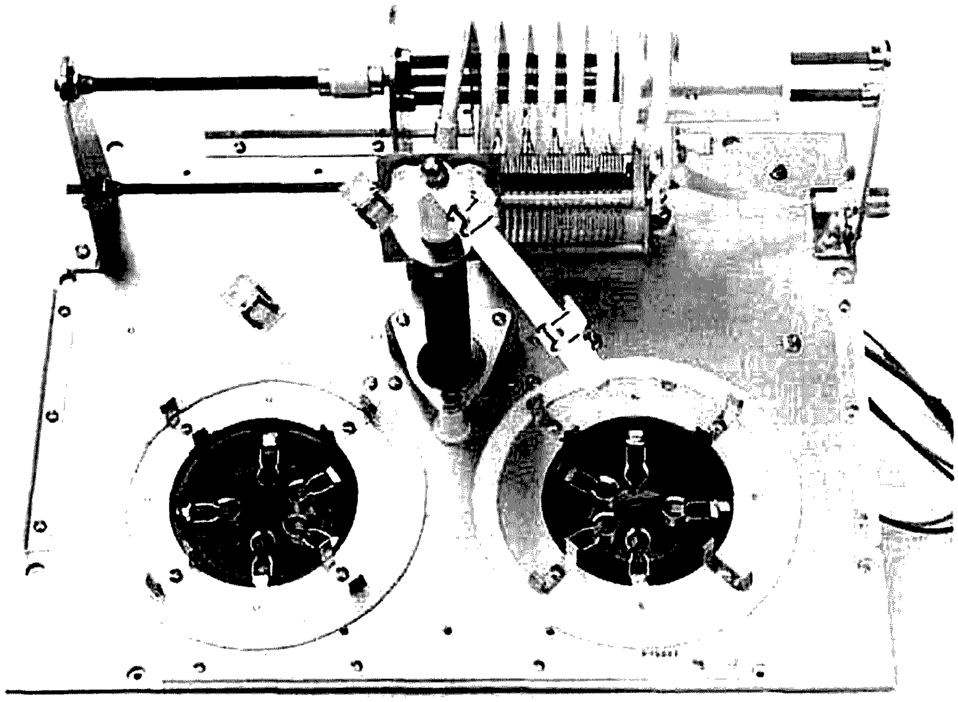
threaded bushing is attached to one disc, through which the drive shaft extends. The shaft is ¼-inch-diameter copper rod, threaded with a ¼-20 die. Two fiberglass guide rods are mounted between the end support brackets to keep the shorted turn from rotating as it's driven back and forth. The threaded drive shaft is driven from the amplifier panel by an insulated extension shaft and an insulated coupling. The

shorted turn is ungrounded at all times, and no moving parts carry rf current.

The plate tank coil is wound with ¼-inch-diameter silver plated copper tubing with an I.D. of 2-¼ inches, providing ample clearance for the shorted turn to move within the coil without danger of arcing. Heavy-duty copper mounting lugs are silver soldered to the coil. Resonance is initially established with a grid-dip oscillator. Indication of resonance is quite broad. Loading is accomplished with the pi network variable output capacitor, C2.

value will cause excessive fundamental rf power to dissipate in the parasitic resistor. The proper value of shunt inductance, while not particularly critical, should be determined for the operational range of the amplifier in each case.

The amplifier shown in the photographs was designed for commercial service and is included in a shielded cabinet as part of a larger package. A similar design using 4-1000A's has been built for commercial ssb service at the 6-kW PEP level—but that's another story. The basic design is



**top view of the high-power commercial amplifier. Inductively tuned tank is shown at rear. Tuning is the same as with a variable capacitor, except inductance decreases as shorted turn penetrates coil. Parasitic suppressors have been removed for this photo.**

## parasitic suppression

An important consideration in plate-circuit design is a parasitic suppressor. In this amplifier parasitic suppressors are included in each plate lead. Each suppressor is a 40-ohm, 16-watt *Glo-bar* resistor shunted across an inductor, which consists of a length of plate lead. The value of the shunt inductance is important. A too-small value won't completely suppress the tendency for vhf parasitics, and a too-large

adaptable to any high-power amplifier operating in the upper region of the hf spectrum.

### references


1. William I. Orr, "Pi and Pi-L Networks for Linear Amplifiers," *ham radio,* November, 1968, page 36.
2. William I. Orr, "The 3-500Z in Amateur Service," *ham radio,* March, 1968, page 56.


ham radio

# a versatile solid-state receiver

The answer to many receiver problems: a circuit featuring high selectivity for cw, ssb, and rtty

George W. Jones, W1PLJ, 12 Trail Street, Cambridge, Massachusetts 02138

With transistors in their present state of development, it hardly seems worthwhile to use vacuum tubes in receiver circuits. At one time it could be argued that cost was a factor, but things have changed. For example, the RCA 40673, a gate-protected dual-gate mosfet, costs about $2.00. In a mixer circuit it will do the same job as a 6BE6, but with fewer components and less power. The mosfet costs a few cents more than the tube, but you can eliminate a screen dropping resistor, screen bypass capacitor, and tube socket, and a big power supply isn't required.

Several mosfet's are available at less than a dollar, such as the RCA 40468 (75c). Very few tubes can be obtained for this amount. For about $4.00 you can buy not one, but *four* IC audio amplifiers in one package: the RCA CA3048. Are you convinced?

## system description

The solid-state receiver described in this article is used with the components shown in **fig. 1**. A recommended converter is described in reference 1. Typical circuits for the vfo, or local oscillator, appear in reference 2 and 3. The 80-meter tuner is described in reference 4.

## features

The solid-state receiver uses mosfet's in rf and mixer stages to reduce cross

modulation. IC's are used in the product detector and audio amplifier. A 9-MHz crystal filter is used for ssb reception. An additional crystal filter, operating at 2.215 MHz, with a 250-Hz bandwidth, provides passband tuning. This allows the i-f passband to be shifted a limited amount without changing the beat note of received signals. Another feature is that the i-f system provides reception over a 2.8-kHz bandwidth while allowing you to boost the one signal you're interested in.

the 2.215-MHz filter. The filter output is then converted up to 9 MHz and applied to the product detector.

The 11.215-MHz oscillator frequency is variable over a small range, so that any slice within the 9-MHz filter bandwidth can be passed through the 2.215-MHz filter. Varying the 11.215-MHz oscillator frequency doesn't change the beat note of any signal, because the same frequency is fed to both mixers. Thus it's easier to keep track of the signal you're trying to receive.

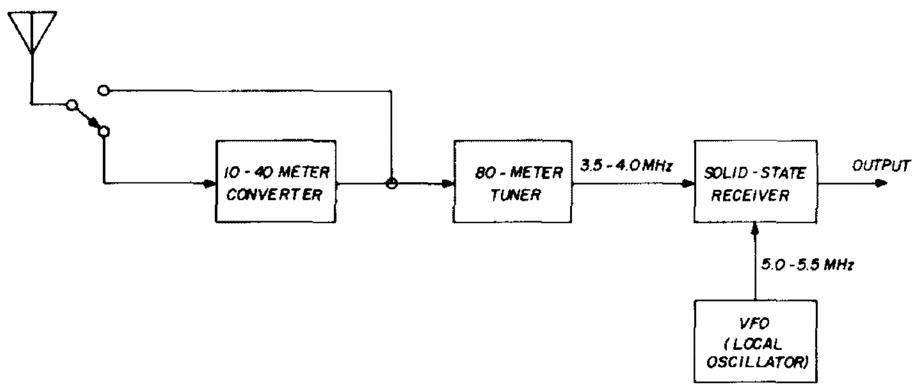



**fig. 1. Integrated receiving system for high performance and versatility. Recommended circuits for the converter, 80-meter tuner, and vfo are given in the references.**

## operation

A block diagram of the solid-state receiver appears in **fig. 2**. Output from the 80-meter tuner, 3.5-4.0 MHz, is mixed with the vfo signal, 5.5-5.0 MHz, to produce a 9-MHz difference signal. This is applied to the crystal filter. If the entire 2.8-kHz bandwidth of this filter is desired, 9-MHz i-f amplifier gain is increased. This allows you to hear what's going on within 2.8 kHz when listening for cw stations, or allows audio components to come through when receiving ssb.

## selectivity

If you wish to listen to only one cw signal in interference, the 9-MHz i-f gain is turned down, and the gain of the two mixers on each side of the 2.215-MHz filter is turned up. The 11.215-MHz oscillator is turned on and tuned so that the desired signal is converted down to a frequency that's within the passband of

## detector

The 9-MHz signal, from either the 9-MHz i-f amplifier or from the second mixer in the 2.215-MHz filter system, or both, now reaches the product detector. Here it's mixed with the output of a crystal oscillator operating at 9.0015, 8.9985, or 8.99745 MHz. The first two frequencies allow sideband selection on ssb, and the third allows standard tones of 2125 and 2975 Hz to be obtained from an RTTY signal centered in the 9-MHz filter's passband. Product detector output is then applied to an IC op amp, in which most of the receiver's gain is produced.

## af amplifier

With most of the system gain developed in the af stage, cross modulation is minimized because the rf signals are at low level in the preceding stages. The af amplifier will drive a pair of crystal headphones; however, a power amplifier

is required if you want to use a speaker.

The af stage drives a one-stage amplifier, whose output is rectified to provide amplified agc. This stage's output also drives an S meter through an emitter follower.

the presence of strong signals. Automatic gain control is also applied to the mixer, and a diode gate passes the appropriate signal to control stage gain.

Mixer output is coupled through a tuned circuit to a McCoy 48B1 crystal

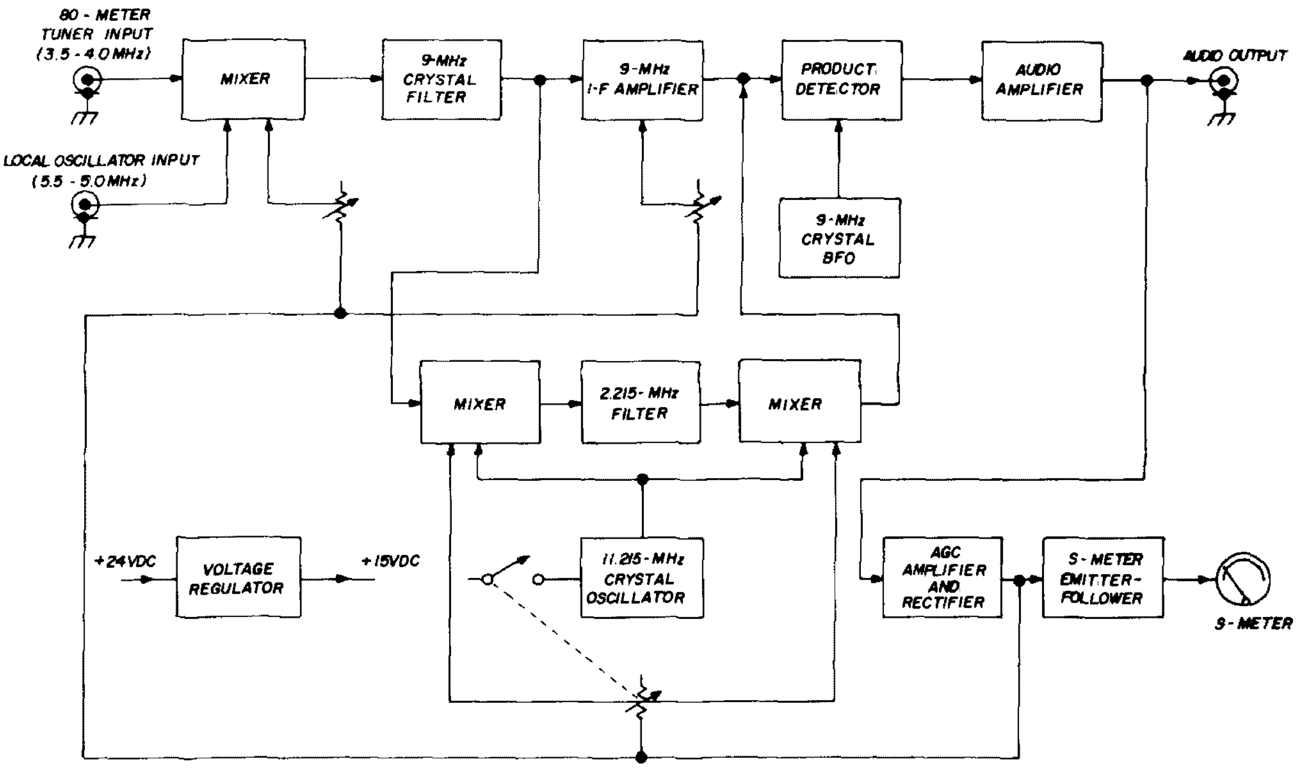



**fig. 2. Functional block diagram. Filter and oscillator frequencies are nominal; see text.**

Power is obtained from unregulated positive and negative 24-volt supplies. A simple 15-volt regulator supplies power for the fet's and ICs.

## the circuits

The first mixer, input tuned circuit, and i-f system are shown in **fig. 3**. An rf amplifier isn't used, because the mixer gives adequate sensitivity for either 80-meter operation direct from the antenna, or from other bands with a converter. Omitting an rf amplifier reduces cross modulation from strong signals by keeping input-signal gain low preceding the filter. The input circuit is from a solid-state receiver that appeared in another article.[5]

## mixers

A manual gain control reduces gain in

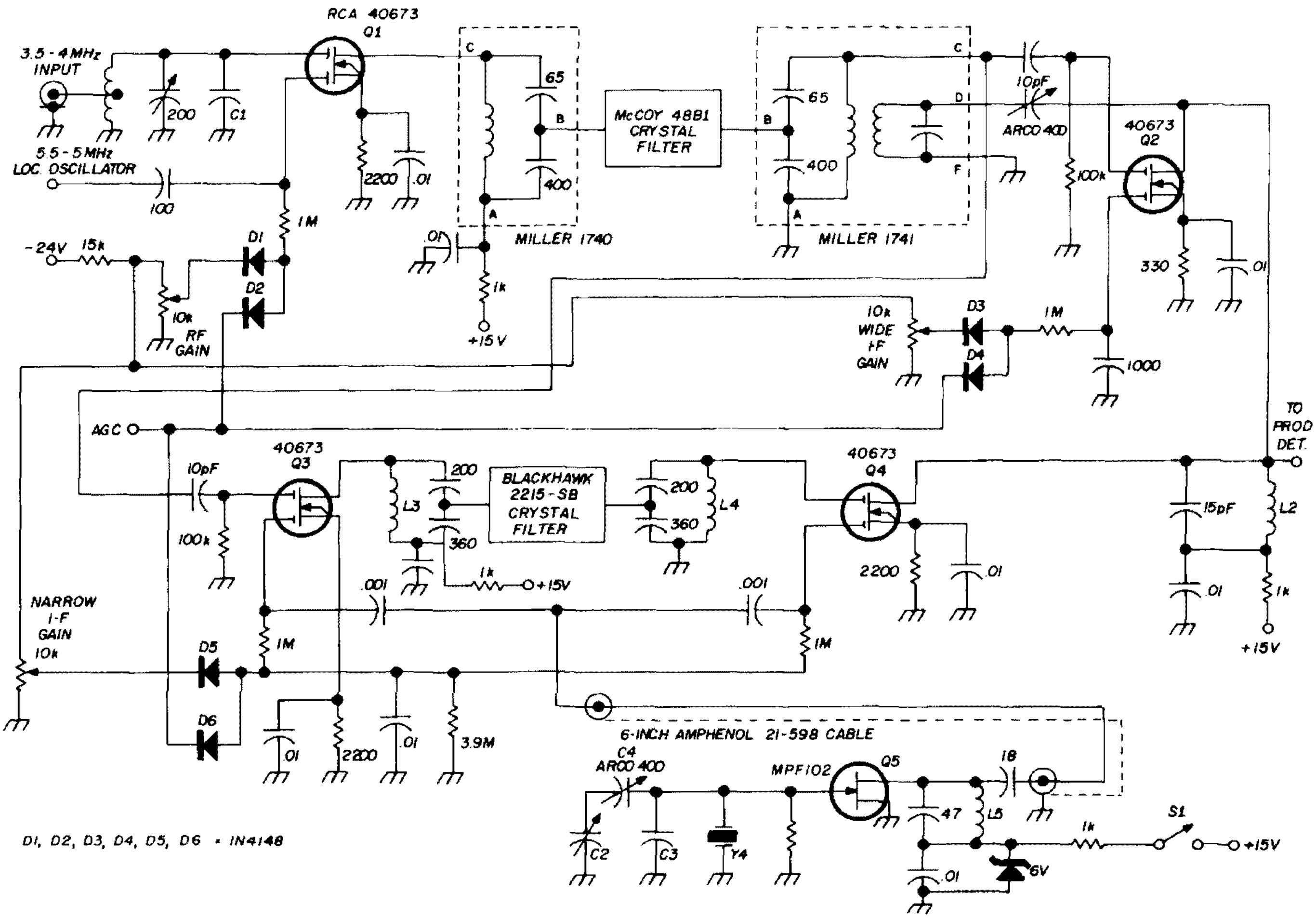
filter. Although the Miller 1740 and 1741 transformers have two windings, I obtained more gain by using only one winding as shown; probably because the windings are loosely coupled. Filter output is coupled to the i-f amplifier, and the otherwise unused output coil winding is used to neutralize the stage. Neutralization is necessary to reduce feedthrough when stage gain is reduced and the narrowband section is used.

Mixer Q3 converts the 9-MHz i-f signal to 2.215 MHz, which is passed through the narrowband filter. Mixer Q4 then converts whatever signal manages to get through the filter to 9 MHz, which is applied to the product detector. Since both mixers use a common oscillator, the frequency that comes out of Q4 is the same as that which went into Q3; there-

fore, if the signal can be heard at all, it's beat note is the same regardless of which i-f went through or the tuning of the 11.215-MHz oscillator.

The tuning of the 11.215-MHz oscillator determines what portion of the wide MHz, because 9002 kHz won't pass through both filters. Therefore, the oscillator must cover 11.2136 to 11.2164 MHz and no more.

A variable capacitor across the crystal varies its frequency over this narrow




| | | | |
|---|---|---|---|
| **C1, C2** | **100 pF (Hammarlund HF-100)** | **L2** | **28 microhenries (CTC 2060-4)** |
| **C3** | **5 pF must be chosen for individual crystal (see Text)** | **L3, L4** | **63 microhenries (CTC 2060-6)** |
| **L1** | **23 turns Airdux 632-T, tap at 1½ turns, 32 tpi, ¾" diameter** | **L5** | **15 turns no. 22 wire on 3/8" diameter slug-tuned form (CTC 1534-2-2)** |
| | | **Y4** | **11.215 MHz crystal** |

**fig. 3. Rf and i-f circuits. Features include low intermodulation products and passband i-f tuning for cw.**

filter's passband goes through the narrow system. To get reasonable sensitivity and make best use of the filters, it's necessary that a slice within the wide filter's passband be converted to the frequency of the narrow filter. The wide filter passes frequencies between 8998.6 and 9001.4 kHz. It will do no good to convert a frequency of, say, 9002 kHz to 2.215 range. C2 is a front-panel control for moving the passband, and C3 and C4 limit the variation so that C2's tuning will be less critical. Some experimentation will be needed to arrive at the correct values for C3 and C4.

The six-inch length of shielded wire could have been avoided with better layout. No trouble has been experienced

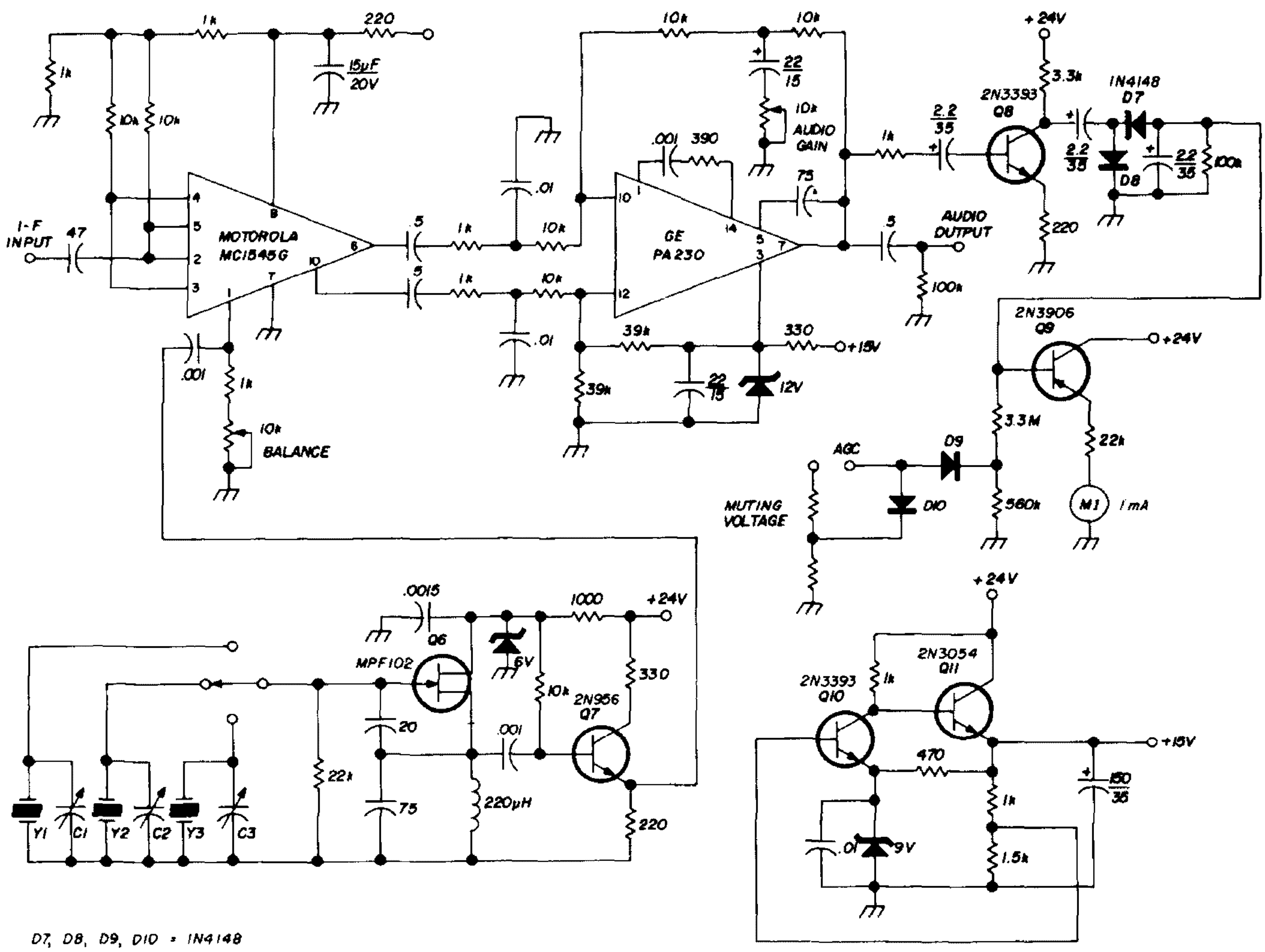



fig. 4. Product detector, af stage, and regulated power supply. The MC1545 acts as a double balanced modulator; output is the sum and difference of input frequencies.

C1, C2, C3 Trimmers (Arco 402)

| | |
|---|---|
| Y1, Y2, | 8.9985- and 9.0015-MHz crystals included with McCoy 48B1 filter |
| Y3 | 8.99745-MHz (or 9.00255-MHz) crystal |

with feedthrough around the filter through the common oscillator connection, nor has the 11.215-MHz oscillator caused any birdies.

## product detector, audio and agc

**Fig. 4** shows the product detector, bfo, audio amplifier, and amplified agc circuits. A Motorola MC1545G is used as a product detector. When wired as shown, it acts as a double-balanced modulator, cancelling in its output any input to pin 1 alone or to pins 2 and 5 in parallel, but giving as output the sum and difference frequencies. The price of the MC1545 is rather high ($8.25), and unless Motorola comes out with a less expensive version, you'll probably want to use another product detector such as one of several suggested in a recent article.[6]

The balance control should be adjusted with an audio signal applied to point A through a capacitor and should be set for minimum audio voltage at pin 6. Product detector output is amplified by a GE PA230 operational amplifier with an audio gain control in the feedback loop. One quarter of an RCA CA3048 could have been used instead with less external components, but with less gain and higher output impedance.

Bfo voltage for the product detector is supplied by crystal oscillator Q6. Y1 and Y2 are used for ssb reception. Y3 is used for RTTY reception with reversal of mark and space *frequencies accomplished in an*

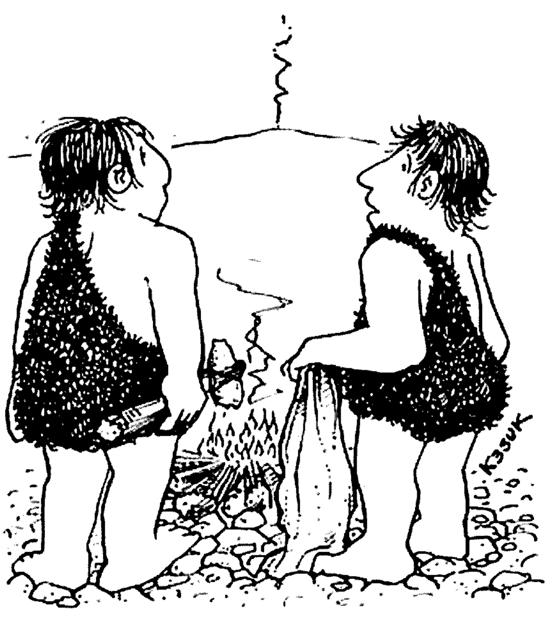
external adapter.

The audio amplifier IC output is applied to agc amplifier, Q8. The agc amplifier is operated without bias so that no output is obtained until the input exceeds 0.6 volt peak, enough to turn on the transistor. Once this threshold is reached, very little additional voltage is needed to get full output from the agc rectifier. This results in a very flat agc characteristic. Some trouble was experienced with motorboating in the agc, which was cured by reducing Q8's gain with the 220-ohm resistor in the emitter lead.

Q9 drives the S meter. S-meter sensitivity can be increased either by decreasing the resistance in series with the meter or by increasing the value of the 3.3-meg resistor, so that more agc voltage must be produced for a given amount of gain reduction.

A diode gate supplies either the agc voltage or a muting voltage from an external source to the agc line. A negative muting voltage is needed while transmitting, and the resistors associated with the muting voltage depend on the voltage available.

Q10 and Q11 comprise a simple regulated 15-volt power supply for the fet's and IC's.



"Ten minutes ago he was 40 over 9 . . . then he went down in the mud."

## performance

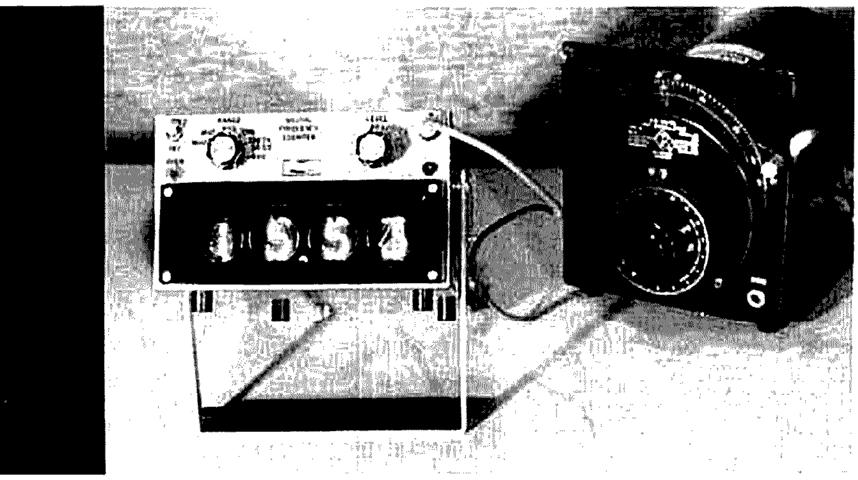
Performance is excellent. Using the wide i-f alone, a signal-to-noise ratio of 20 dB was obtained with 1-microvolt input from a Knight KG-686 signal generator. Using the narrow i-f alone, a 16-dB signal-to-noise ratio was obtained with the attenuator on the signal generator at zero. Avc threshold is reached with 10 microvolts input, at which point output is 0.95 volt for a typical setting of controls. Increasing the input to 100 millivolts, an 80-dB increase, only raises the output to 2.1 volts, a 7-dB increase.

Future plans for the receiver include a general-coverage converter using a phase-locked oscillator so that any multiple of 500 kHz can be used as a mixer frequency, with a panadaptor operating from the output of the first mixer or possibly from a separate first mixer.

### references


1. M. Goldstein, VE3GFN, "Bandswitching FET Converters," *ham radio*, July, 1968, p. 6.
2. J. Fisk, W1DTY, "Stable Transistor VFO's," *ham radio*, June, 1968, p. 14.
3. G. Hanchett, W2YM, "The Field-Effect Transistor as a Stable VFO Element," *QST*, December, 1966, p. 11.
4. M. Goldstein, VE3GFN, and G. Cousins, VE1TG, "High-Quality Hybrid Receiver," *73*, February, 1968, p. 42.
5. G. Daughters, WB6AIG, et al., "Solid-State Receiver Design with the MOS Transistor," *QST*, April, 1967, p. 11, and May, 1967, p. 22.
6. D. DeMaw, W1CER, "Some Notes on Solid-State Product Detectors," *QST*, April, 1969, p. 30.


ham radio



# compact frequency counter

A solid-state instrument featuring direct numerical readout to eight digits with four readout tubes

Bert Kelley, K4EEU, 2307 South Clark Avenue, Tampa, Florida 33609

This digital frequency counter is an improved version of a design that appeared in an earlier issue of *ham radio* magazine.[1] The new design features direct numerical readout (using Nixie tubes), smaller size, simpler construction, and some optional circuits for increased versatility.

Two counter modules are presented. One, using five IC flip-flops, has a nominal upper counting range of 16-17 MHz. The other has a nominal upper range of 12-13 MHz, but uses only two ICs. The simpler construction and lower cost of the latter module should appeal to those with measurement requirements limited to this lower range.

An optional overrange indicator provides a visual warning when the count is higher than that indicated by the readout tubes. Switching logic is designed so that frequencies with as many as eight digits can be displayed using only four tubes.

With many new ICs appearing on the market, the Motorola devices used in this instrument continue to be the lowest priced, and their performance exceeds published specifications. They are truly a best buy for the amateur interested in building fine test equipment at lowest cost.

The B-6091 Nixie readout tubes were a lucky surplus buy; however, other types of readout tubes can be used as long as

they're not biquinary. Be sure to obtain a data sheet with current-limiting resistor values and pin connections. It shouldn't be necessary to increase supply voltage above 170 volts, nor is it recommended.

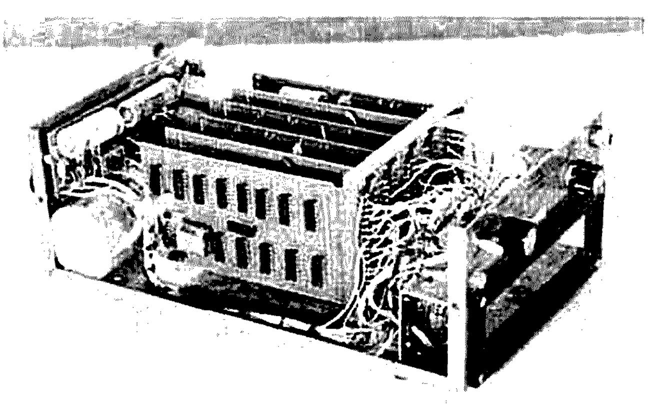
In the design of this counter, I used information from Motorola applications bulletins, experience from previous designs,[1] and suggestions from other amateurs. W2FLJ suggested the 4-MHz clock, and W4IGX helped with the mechanical layout.

A block diagram of the basic instrument appears in fig. 1. All modules including the power supply are constructed on PC boards.* The entire unit is enclosed in a 4x7x12-inch Minibox.

## clock module

This circuit (fig. 2) determines overall system accuracy. A highly stable crystal, as described by Hoff,[2] is used as a standard. The basic frequency is 4 MHz, which is divided twice to 1 MHz, then divided down by a series of decade dividers to 1 Hz. Five different time bases are used. For each time base, the clock generates a short pulse timed to occur prior to the negative-going transition of the main clock pulse. These advanced pulses, as well as the clock pulses, are selected by a range switch (fig. 7) and routed to the control module (fig. 5) for processing and use. Range switch contacts also ground NE-2 "decimal-point" bulbs positioned between the Nixies.

The crystal, an International HA-1, provides frequency stability without an oven. The circuit will oscillate to 10 MHz and will remain within 1 Hz of WWV at 10 MHz if the specified crystal is used and the simple instructions in Hoff's article[2] concerning temperature compensation are followed.



**Interior view of the frequency counter showing the clock module crystal and zeroing capacitor.**

If you plan to build this counter, a 100-kHz crystal should not be used. The inherent accuracy of digital counters is determined by the quality of crystal time base. Most 100-kHz crystals will drift to the extent that at 10 MHz or so readings will be inaccurate, since frequency drift is multiplied by a factor of 100. Time bases using the 60-Hz line frequency have similar problems, and in addition don't lend themselves to the range switching system used in this instrument.

## 17-MHz counter module

Note that in the beginning of this article I mentioned that the *nominal* upper counting range of the counter module was 16-17 MHz. The maximum upper counting range depends on several factors: the JK flip-flops, skill of the builder, and even luck to a certain extent. If reasonable care is used in construction, the module should count to 17 MHz.

This module (fig. 3) uses five flip-flops in a ring counter, as described in reference 3. The ring counter is decoded by three MC724P dual quad input gates that drive 2N1893 high-voltage transistors. These ground the appropriate pin of the readout tubes. An MC799P satisfies loading and polarity requirements.

When completed, the counter should be checked with a conditioned pulse, such as the 1 Hz output from the clock. It probably will be necessary to reset the JKs with a 4-V pulse on the reset bus (by briefly shorting pins 1 and 2 of the counter decade under test). With all voltages applied, the Nixies should advance through 1-2-3 etc., to zero and start

*Printed circuit boards are available from Stafford Electronics, 427 So. Benbow Road, Greensboro, N. C. 27401. Kits, including boards and all components, are also available. A set of full-size layout drawings for the boards is available from *ham radio* magazine for 75¢.

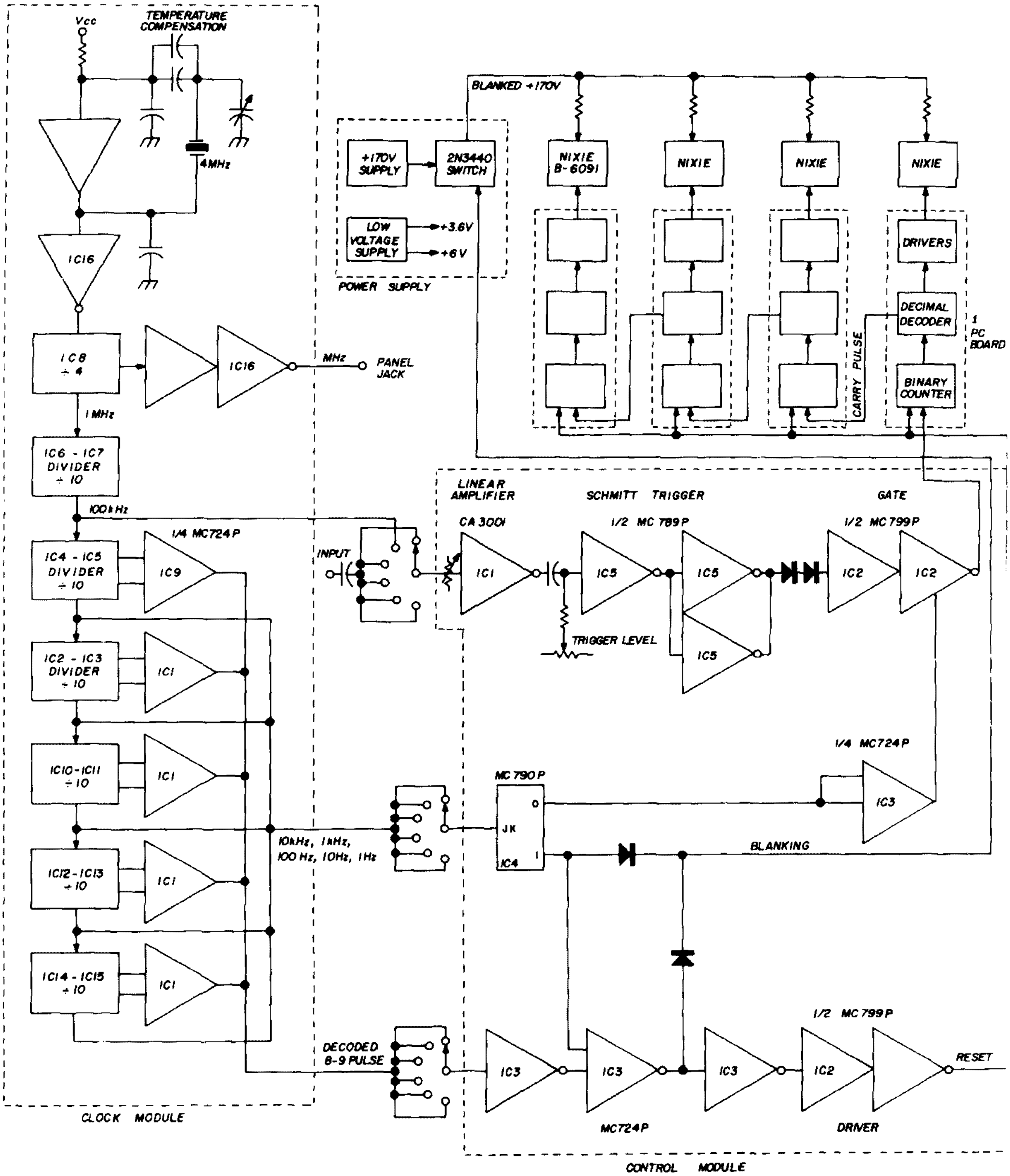



**fig. 1. Block diagram of basic instrument. The "decoded 8-9 pulse" is an advance pulse that resets the counte**

over again.

If the decade hangs up on a number, or more than one number lights, the work starts. Reference to the IC truth tables and other published information[4] should help to locate the defect.

The counter module shown in **fig. 4** uses two of the latest Motorola RTL ICs, the MC780P flip-flop and its companion decoder, the MC770P.* The MC780P contains four internally connected flip-flops, with four outputs for decoding and the usual toggle and reset inputs. In

*Listed in Allied's industrial catalogue supplement 692. The MC780P and MC770P sell for $4.05 and $3.55 respectively.

addition, four direct set inputs can be used to set the IC to any status—a feature not needed for a frequency counter, but useful for digital clocks.

Since this combination simplified construction and reduced cost, I obtained a supply of these ICs and built decades to check their performance. I found the MC780/770P combination was not, and performance might vary slightly with other units.

I made some checks to determine if decade/decoder combinations could be intermixed within the same counter module. I found them to be incompatible because of different carry-pulse polarities.

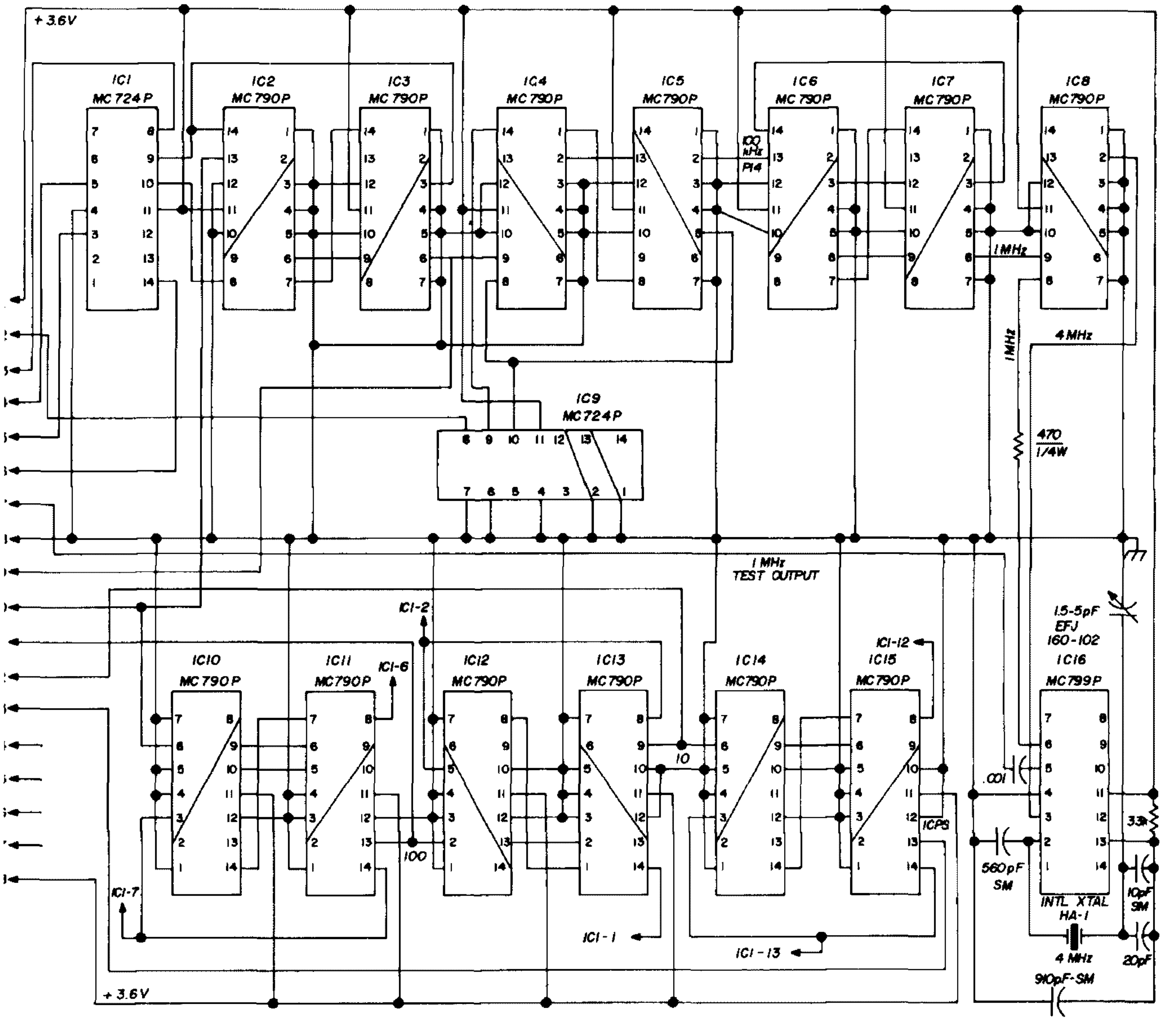



**fig. 2. Clock module. Circuit provides outstanding frequency stability without an oven.**

maximum reliable count frequency to be between 12 and 13 MHz. The more complex ring counter consistently counted to 16 or 17 MHz without special selection of components. Although the ring counter was checked with several decade-decoder combinations, the However, they operated the overrange indicator properly, and no problems should occur if four similar decade/decoder combinations are used in the same counter module.

If you are satisfied with a top frequency of 12 MHz, this new counter design

with the two special ICs will greatly simplify construction and reduce cost.

## control module

A linear integrated circuit amplifier, the CA-3001, results in a simple control module design (fig. **5**). The CA-3001 has good response to at least 20 MHz. It has two outputs, only one of which is used, and a built-in gain control provision at pin 2, which offers a smooth, non-frequency-sensitive method of controlling circuit gain. I tried this method of controlling level, but the direct potentiometer approach offered better protection for high-input levels.

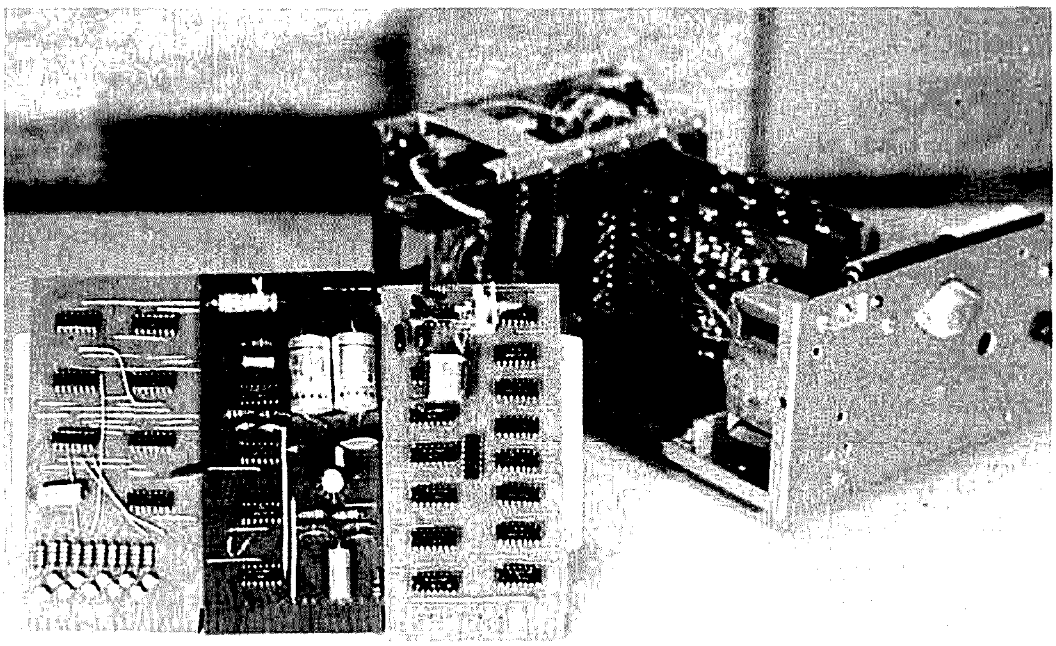
The MC789P Schmitt trigger shapes and limits the incoming signal, and diodes D3, D4 translate the dc level between MC789P and MC799P, acting as a form of zener. Junkbox silicon diodes may be used if they pass a simple test: connect the diode under test in series with a 1k resistor and monitor the voltage across the diode while the output from a power supply across the combination is varied from 0 to 10V. The diode voltage drop should be near 0.65V. Unsatisfactory diodes will vary widely. A similar test may be made for D1 and D2 by testing for a junction drop of 0.25V. Many diodes won't pass this test, and many junkbox diodes assumed to be germanium will turn out to be silicon. Of course, new diodes obtained from a reputable manufacturer won't require this test.

## logic and timing

Referring to figs. **1** and **5**, IC2 operates as a combination driver and gate by using IC3-14 as a switch to pull an internal element of IC2 to ground when inhibiting the count. If an unprocessed count pulse were applied to pins 13, 14 of IC3, the gate would open for only half the required time.

Pulses from the tens decades in the clock module are not symmetrical. IC4 restores pulse symmetry and delivers two voltages of opposite polarity and equal amplitude.

If a 1-Hz gate time is selected by the range switch, the following events occur (see the timing diagram, fig. **6**): the



**Printed circuit boards used in the counter: left to right, counter module, control module and clock module. Four counter modules in the counter are identical.**

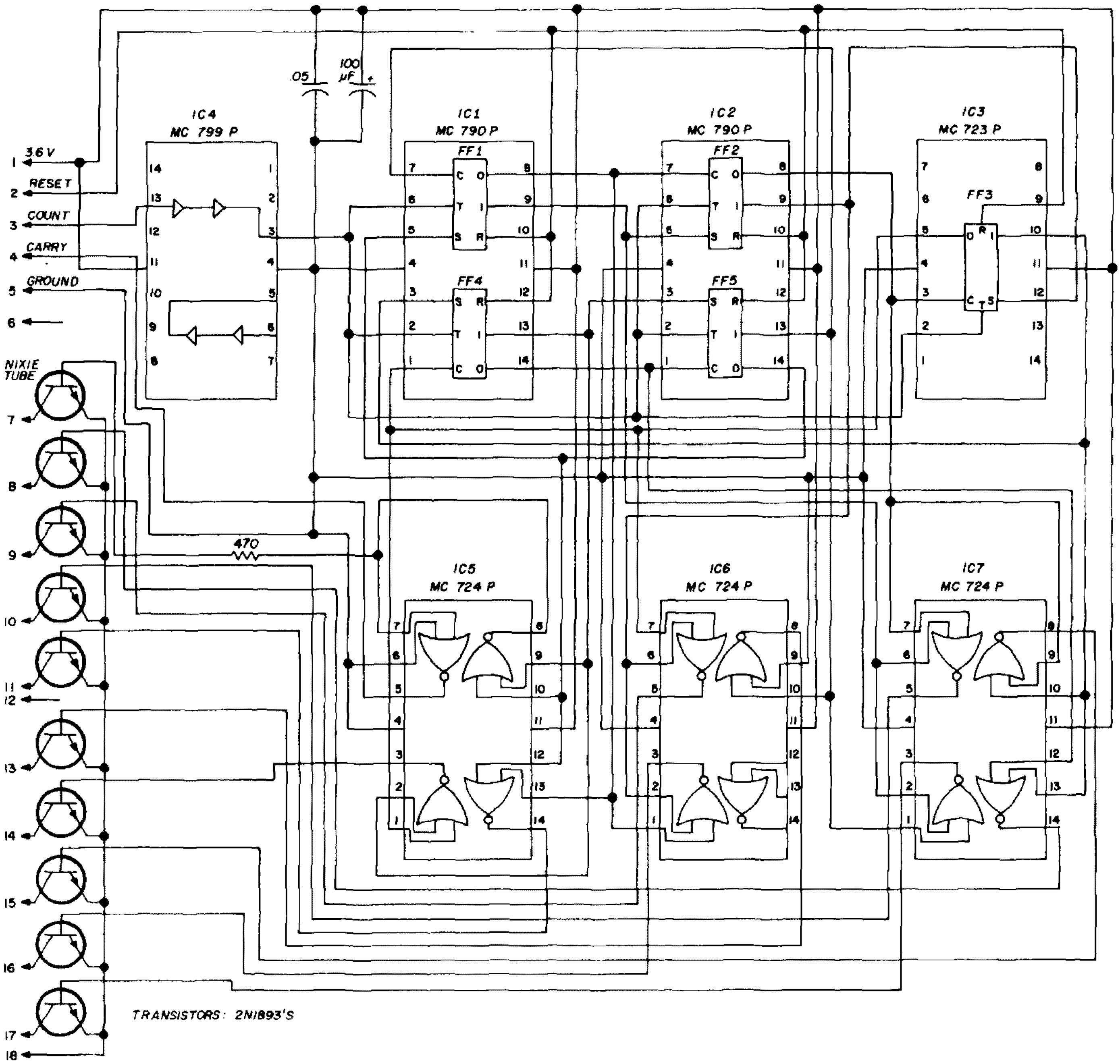



**fig. 3. Basic counter module. Ring circuit provides nominal upper counting range of 17 MHz.**

output from IC4 occurs at a 0.5-Hz rate. IC4-8 is positive for one second; IC4-9 is positive for one second, then its output falls to zero, allowing a count pulse to pass through IC2-5.

## to the counter decades

If IC4-9 output is low, IC4-8 is positive approximately 1.4 V, saturating one-half of a gate section connected to IC3-7. Diode D1 then conducts, which places a positive voltage on the blanking line at pin 15 on the module and the power supply, where it shuts off the 2N3440 high-voltage switch that controls power for the readout tubes.

If IC3-7 is saturated, the gate won't pass the advanced clock pulse; but when the next negative-going transition of the clock arrives, the JK reverses state and shuts off the count: the blanking voltage via D1 goes to zero, and the readout tubes light, displaying the count. At the same time the voltage will have dropped on IC3-7, and it will be ready to pass the next advanced pulse, which occurs near the end of the display interval.

When the advanced pulse arrives, it is

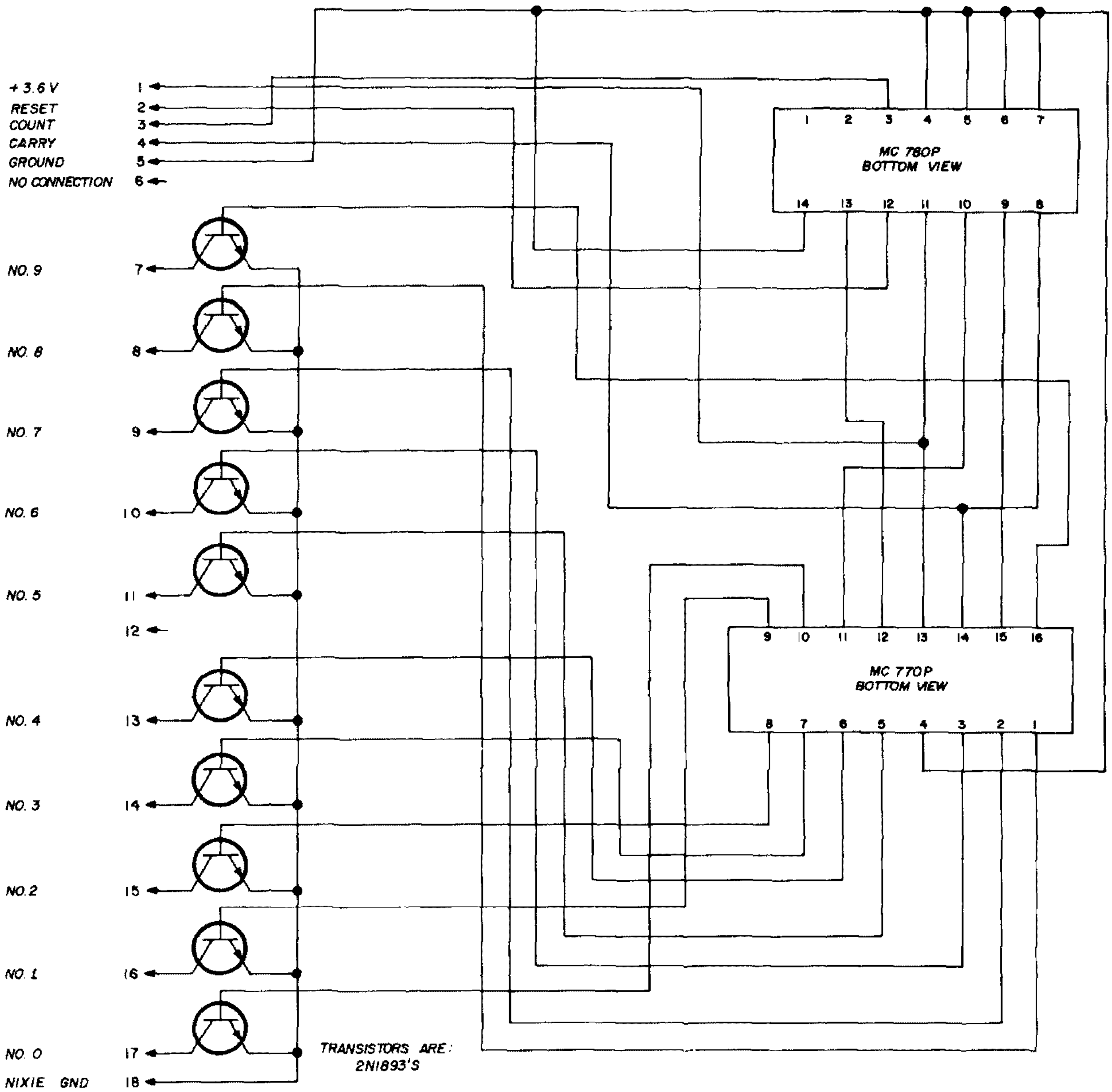


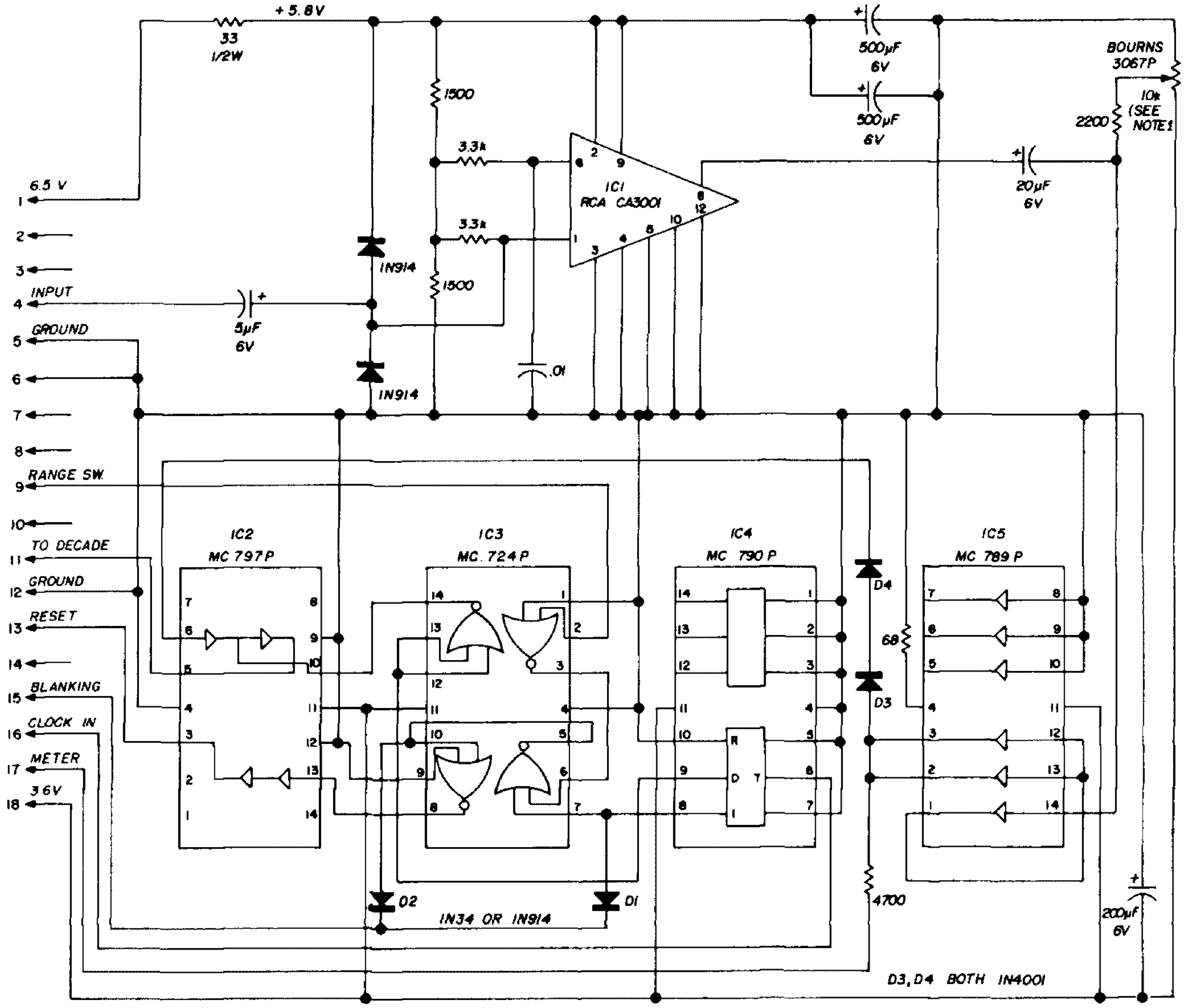

**fig. 4. Alternate counter module using IC's that became available after completion of basic circuit.**

inverted, gated, and allowed to reset the counters to zero. D2 conducts and shuts off the Nixie power so that when the next clock pulse arrives, the counters are cleared, the Nixies are off, and the JK changes state. D1 picks up the positive voltage on IC4-8 so that blanking remains off during the new count interval. This entire process takes place at speeds from 1 to 10,000 times per second, depending on the setting of the range switch.

Some blinking will occur at the two lowest speed-count settings, but the display will be continuous at the other three speeds. It is possible to eliminate the blinking by the use of storage. Blinking is not objectionable; therefore storage was not used.

## overrange display

It is possible to measure a signal with a frequency much greater than four digits by sampling the incoming signal at a fast rate. As mentioned previously, gate times as fast as 10,000 counts per second are used. If the incoming signal happens to be, say, 11,234,200 Hz and the gate is open at the fastest rate, then 1123 would register on the readout tubes—the last four digits would be dropped. If "decimal points" are connected to proper contacts on the range switch **(fig. 7)**, the reading




NOTES:

1. PRESET AT MIDRANGE. AT FINAL ADJUSTMENT, ALTERNATELY ADJUST INPUT LEVEL AND THE 10k POT FOR MAXIMUM (15 TO 17 MHz), THEN LEAVE AT THAT POSITION.

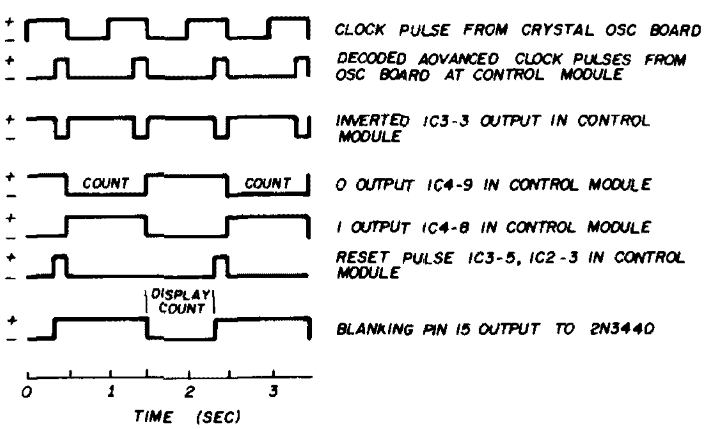
2. OBSERVE DIODE POLARITY. THE END WITH THE BANDS IS THE CATHODE.

**fig. 5. Control module. The CA-3001 input amplifier has a response to 29 MHz.**

will be directly converted to MHz or kHz, eliminating mental arithmetic.

Now, if the range switch is rotated to the 1-Hz sampling rate, the remaining

**fig. 6. Timing of control-module pulses.**




four digits will be indicated, with the counter now reading 4200. (The excess count wduld have spilled out of the left hand decade in the form of carry pulses, lighting the overrange indicator.)

In counting and resetting, all digits on each Nixie are activated, but are permitted to light only during the display period by a blanking circuit, which rapidly switches the power to the Nixies. Thus, the readout tubes are strobed at the same frequency as the gate, and the count appears to stand still.

The overrange indicator (fig. 8) activates a neon light when the count is higher than that indicated by the readout tubes. The MC723P toggles on the first

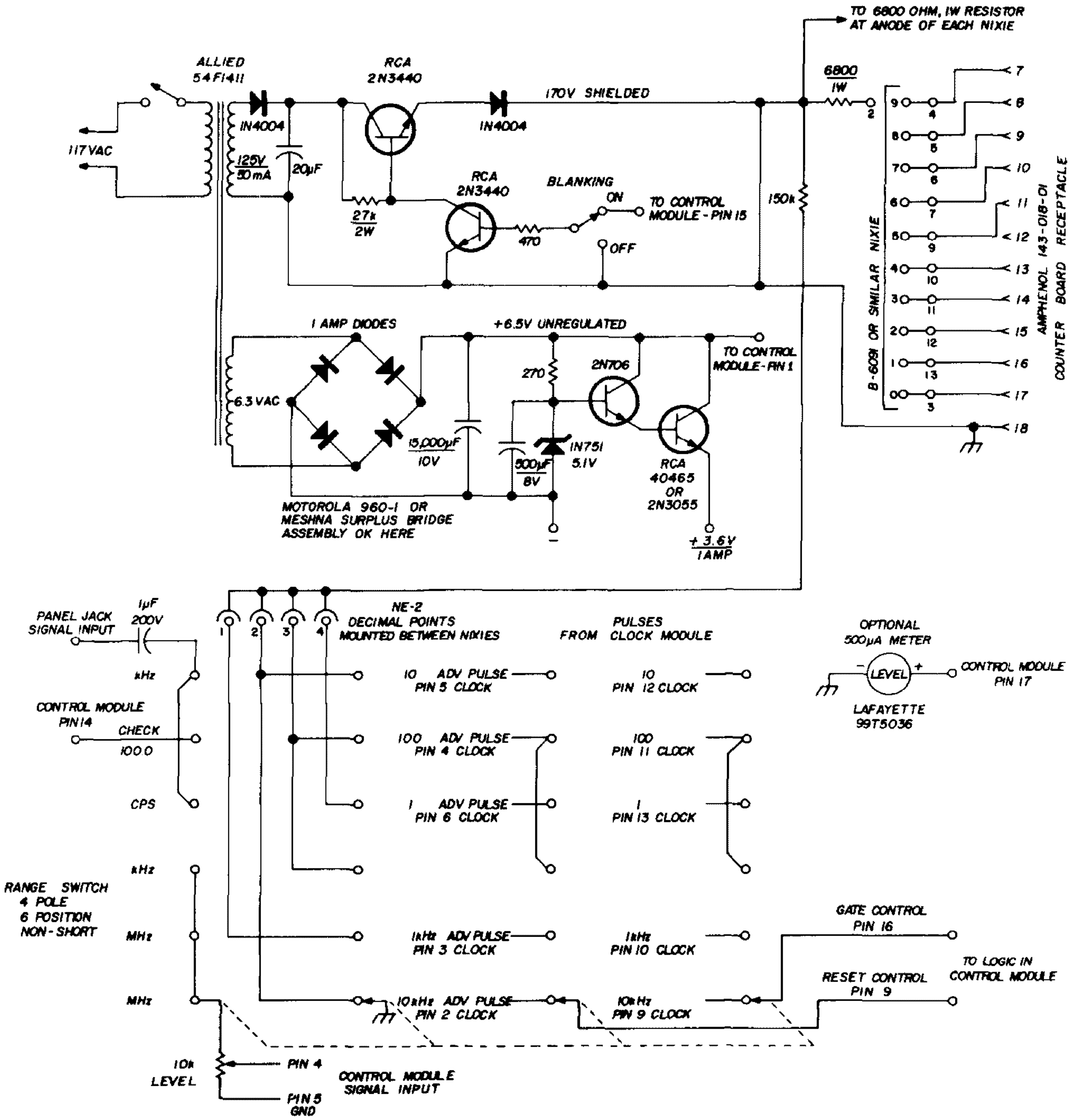



**fig. 7. Power supply and range-switch schematic.**

carry pulse appearing at the left-hand counter decade and locks up until the next reset pulse.

## power supply

The circuit shown in **fig. 7** is a dual-purpose power supply that furnishes all operating voltages for the counter. An amplifier-switch, consisting of an emitter follower and amplifier, provides high voltage for the readout tubes. Unregulated 6.5 volts for the IC amplifier in the control module is taken from the top of the low-voltage supply. A Darlington pair-zener configuration provides regulated 3.6 volts for the flip-flops.

## construction notes

A low-wattage soldering iron is essential when working with solid-state components. The inevitable debugging in a project such as this will be simplified if you have a vacuum-operated de-solder device, such as one of those made by Soldapult or Ungar.

The power-supply board is mounted

on ½-inch spacers on the rear inside wall of the minibox. The large filter capacitor, power transformer, and 2N3055 power transistor should be mounted so that the minibox cover will fit.

Negative leads should not be grounded at the power supply, but should be routed to a common point near the front of the instrument. The 170-volt supply ground return is via the outside conductor of a length of miniature coax cable; the inner conductor carries the high voltage. All power leads are in one cable, which runs along the left side of the instrument.

A standard Minibox results in a compact instrument that fits nicely on a crowded workbench. A ½-inch aluminum strap was bent to form a handle, which also serves as a prop to position the instrument for easy viewing.

Readout tubes were mounted on an aluminum bracket directly behind the PC board sockets. After the Nixie-tube sockets were mounted, masking tape was used to cover the sockets and the inside of the bracket. The corresponding part of the enclosure was sprayed with dull black lacquer to suppress reflections.

A panel bezel was sawed out of scrap ¼-inch plexiglass, filed to shape, painted, and mounted over a cutout on the end of the minibox. A small glass window for the readout tubes was secured by clips. Datak dry decal transfers were used to mark the panel. The printed circuit boards were mounted on short lengths of ½-inch aluminum angle.

## accuracy checks

Since the input frequency is rarely in phase with the counter gate, the displayed count varies according to the number of pulses in the gate at the time it operates. It is therefore normal for the counter to have a 1-digit error, which extends throughout its frequency range of 20 Hz to 15 MHz. At the higher frequencies, accuracy depends on the crystal clock more than any other factor; but at low frequencies, the error remains within 1 digit. One of the positions of the range switch is used to check the internal circuits of the counter. It may be, when the counter is completed, that the 100.0 test position displays 100.1. If this happens, change the 100-kHz jumper in the clock module from pin 13 of IC6 to pin 14.

The uses of this counter are not limited to its basic frequency. I have set 150-MHz communications monitors on frequency, using the 10th harmonic of a 15-MHz oscillator. W4BNE regularly checks the frequency of a MARS net on 50 MHz using a similar method.

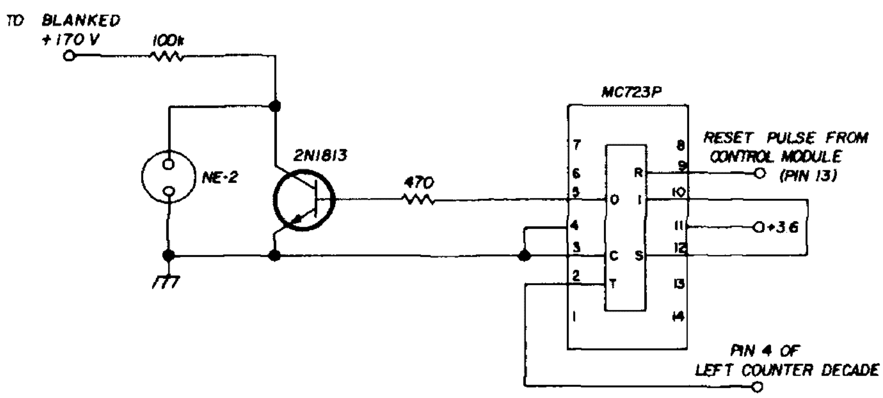
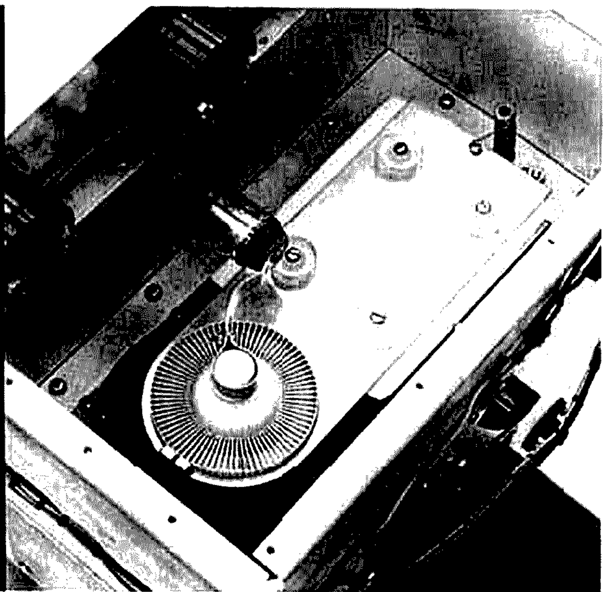



**fig. 8. Overrange indicator schematic. Neon bulb illuminates when fourth counter decade put out a carry pulse.**

### references

1. Bert Kelley, K4EEU, "Digital Frequency Counter," *ham radio,* December, 1968, p. 8.
2. Irvin M. Hoff, W6FFC, "Mainline FS-1 Frequency Standard," *QST,* November, 1968, p. 34.
3. "Application Note AN-251," Motorola Semiconductor Products, Inc., Technical Information Center, P.O. Box 20912, Phoenix, Arizona 85036.
4. Steinback, "IC Counters and Dividers," *Electronics World,* January, 1968.

**ham radio**



# low-drive kilowatt linear for two meters

The Eimac 5CX1500A pentode in a grounded-cathode amplifier using strip line construction

Irwin R. Wolfe, W6HHN, 3467 Rambow Drive, Palo Alto, California 94306

Recent articles in *ham radio* of vhf power amplifiers using strip line tank circuits[1,2] inspired me to boost the output of my Communicator IV to a kilowatt on 2 meters. I used an Eimac 5CX1500A pentode in a grounded-cathode linear amplifier circuit (fig. 1). No drive power is required, which eliminates the need for a buffer amplifier.

## strip line plate circuit

Details of the "sandwich" tank circuit are given in fig. 2. The grounded lower plate is slightly wider than the upper plate to accommodate the antenna loading clip. Four 10-32 screws and nuts and eight polystyrene washers hold the assembly together. Two sheets of Teflon, 15 mils thick, are used for insulation between plates. The capacitance of this sandwich measures 285 pF.

Upper and lower plates are made of 1/8-inch-thick silver-plated copper. The upper plate is cut to accommodate the tube's 3-3/8 inch-diameter cooling fins. The narrow section is cut through, and both ends are bent up and drilled for a 6-32 screw and nut. This clamps the end of the tank circuit to the tube. The lower plate is bent to form the rear support for the assembly. It is grounded by screws at the end farthest from the tube.

High voltage is fed to the plate through an rf choke at the low-impedance point of the upper plate. I used a 3-30 pF vacuum variable for plate tuning, but a capacitor with range of about 2-10 pF should be adequate to tune 144-148 MHz.

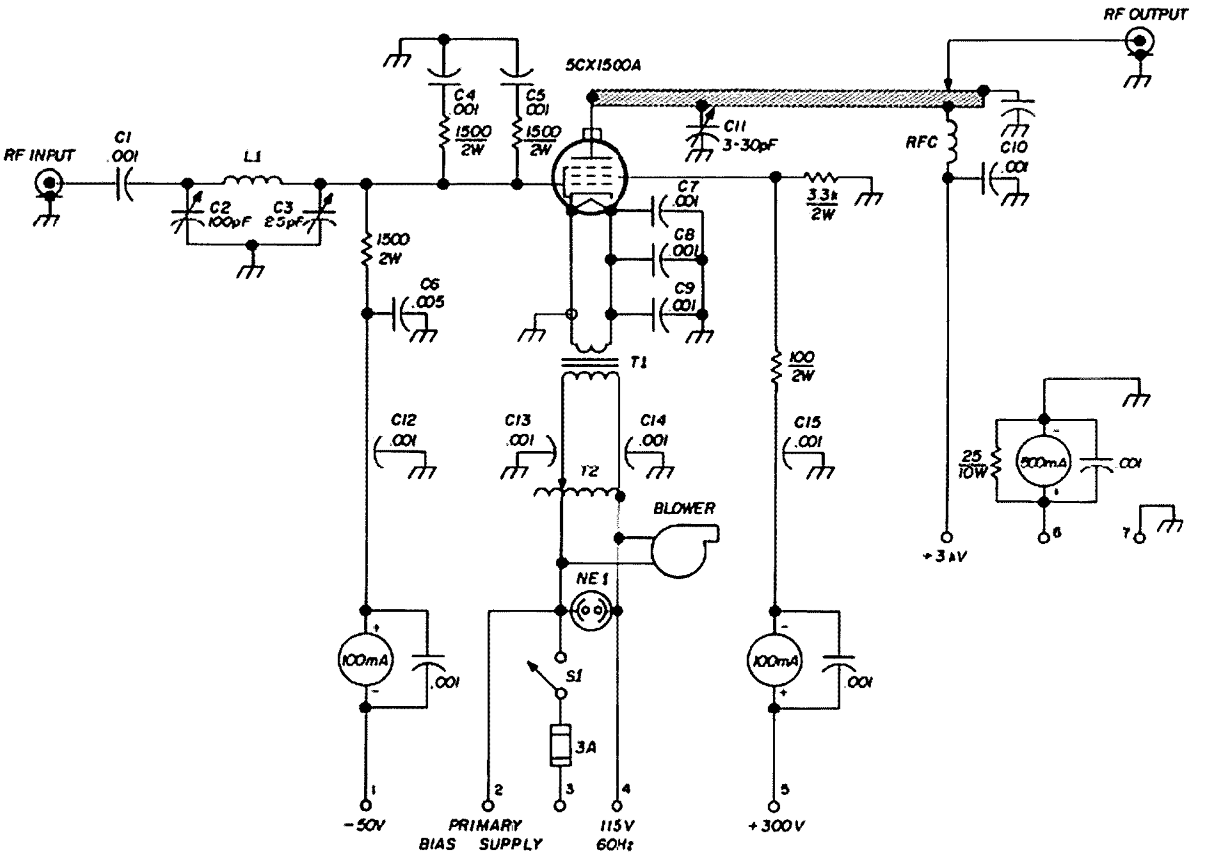
## power supplies

Plate and screen supplies are conventional (fig. 3). The bias supply uses a small line isolation transformer and a silicon-diode bridge rectifier. I couldn't find a 5-volt, 40-ampere filament transformer in any catalog or ham equipment supply store, so I used a type F530 Stancor transformer (5 V, 30 A) with an autotransformer in the primary to obtain 5 volts at the tube-socket terminals. After two hours of operation, the transformer

temperature rise was insignificant.

## enclosure

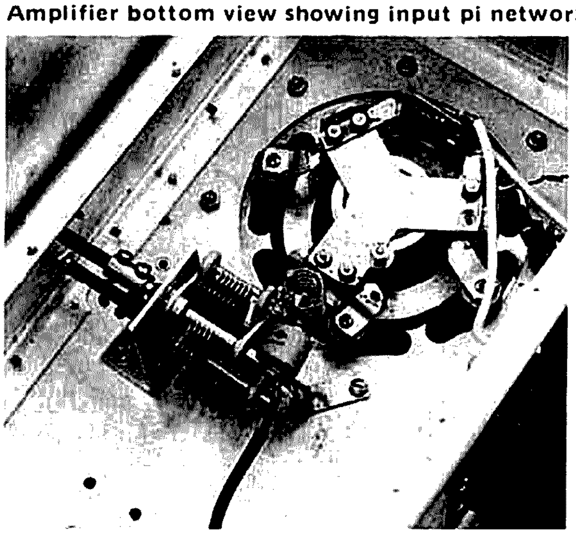
The amplifier is built on a standard 19-inch panel, which is 19¼ inches high to accommodate blower height. An aluminum box (fig. 4) houses the amplifier. A shelf separates input and output circuits. A 3½-inch cutout in the top plate, directly over the tube, permits cooling air to vent out of the cabinet. A perforated metal plate covers the blower cutout. An




| | | | |
|---|---|---|---|
| C1,C7,C8, C9, C10 | .001 µF (Centralab type 858S) | L1 | 5 turns no. 16 tinned copper, 5/8" diameter, 1-1/8" long |
| C12-C15 | .001 µF feedthrough capacitors (Erie type Ck70AW-102M) | Blower must deliver 60 to 100 cfm | |

**fig. 1. Schematic of the easy-to-drive 2-meter linear amplifier. Low interelectrode capacitance of the 5CX1500A eliminates neutralization. Third-order intermodulation distortion is —36 dB.**

**Amplifier bottom view showing input pi network.**



Eimac SK806 chimney confines air flow through the tube cooling fins. The filament transformer primary connections, grid-bias connections, and screen connections are brought out of the box with feed-through capacitors.

The upper control in the front-panel photograph is for plate tuning; the two lower controls are for the input circuit pi network capacitors.

## tuning linearity adjustment

Control of rf drive is necessary to adjust the amplifier for linear operation. I

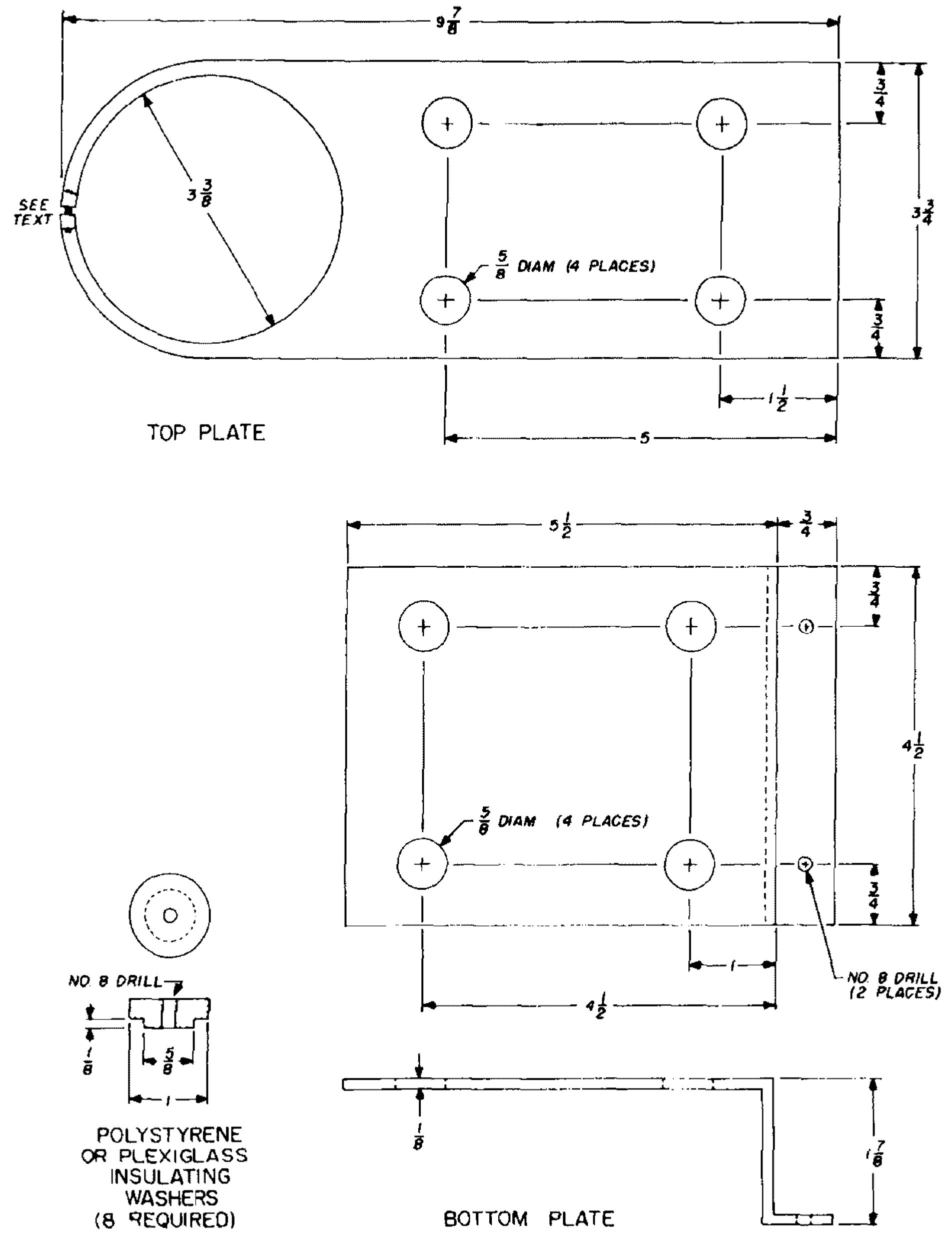
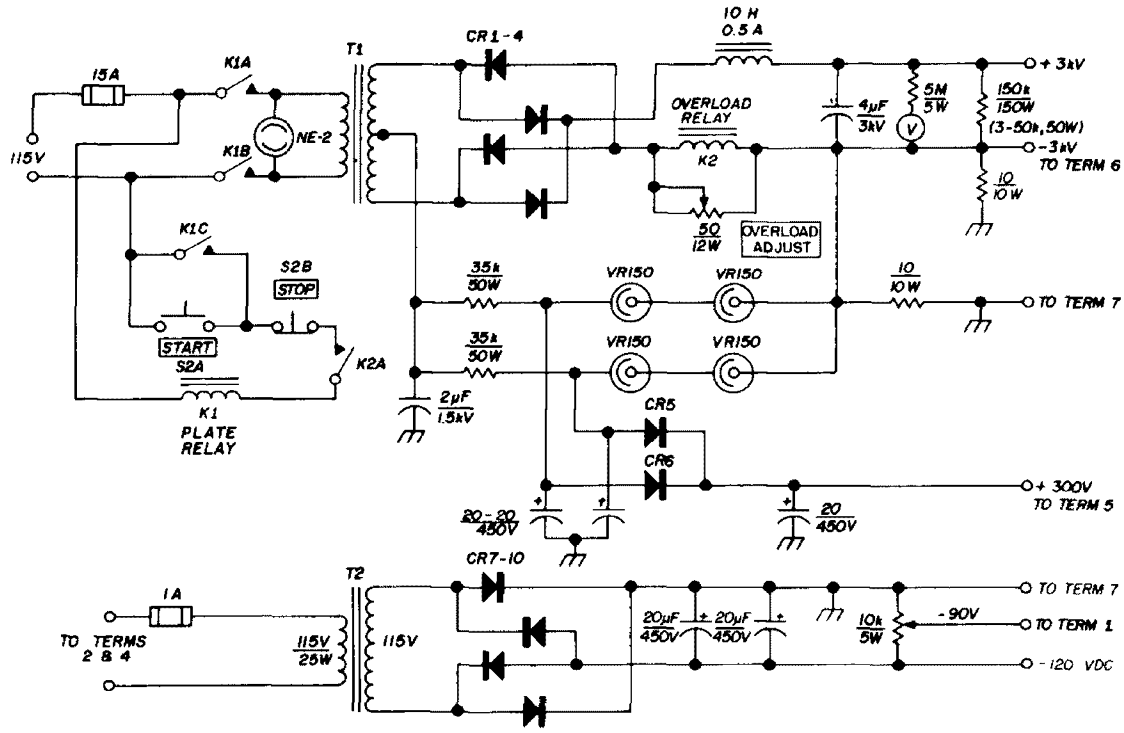



**fig. 2. Construction details of the strip line tank. Material is 1/8-inch silver-plated copper separated by two 15-mil-thick Teflon sheets.**

installed a potentiometer in the driver to adjust the output tubes's screen voltage. To adjust amplifier linearity, proceed as follows.

Connect the exciter output to the amplifier input through a directional coupler. Apply bias and filament power to the amplifier. Adjust both driver tuning and the amplifier pi network capacitors for maximum incident output through the coupler. Switch the coupler to reflected power and adjust the pi network and exciter tuning for a minimum, which should be almost zero. Increase grid bias to maximum, and reduce rf drive. Connect a dummy load and an rf indicator to the amplifier. Place the antenna clip along the bottom plate at a point about 3 inches from the ground end.

Apply plate and screen power, and increase excitation until the output indicator shows power. Tune the plate capacitor for maximum rf output. Do not allow amplifier screen current to increase above 50 mA during this adjustment. Adjust drive and bias for a *plate current* of 350 mA. When the plate tuning capacitor is adjusted, screen current can also be used to indicate resonance.




| | | | |
|---|---|---|---|
| CR1-CR4 | Each diode consists of 8 series-connected 600 V, 1 amp silicon diodes | T1 | 3000 volt center-tapped transformer |
| CR5-CR10 | 600 V, 1 amp silicon diodes | T2 | 117—117 Vac isolation transformer |

**fig. 3. Plate, screen and bias supplies.**

Linearity adjustment should be made with an oscilloscope. I used a parallel resonant circuit to couple the amplifier output to the scope. Complete details for obtaining linear operation are given in reference 3.

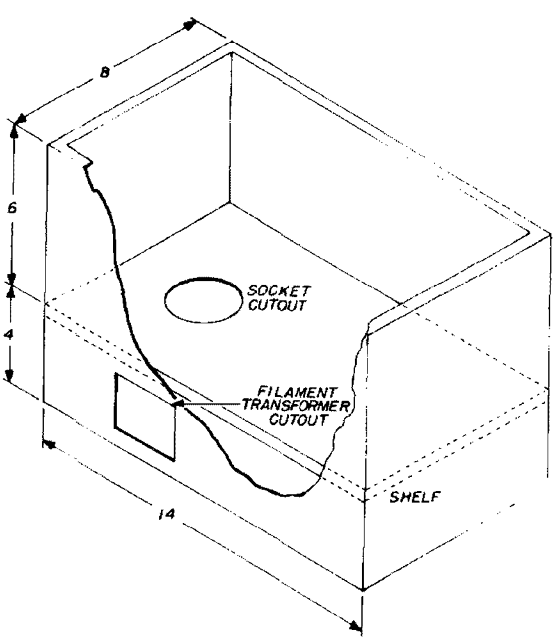



**fig. 4. Rf enclosure detail. Material is 1/16-inch aluminum.**

Loading adjustments are made by moving the clip lead to different positions along the bottom plate of the strip line tank. Make certain that both plate and screen power are removed before making these adjustments. Replace the top plate after each new position, and resonate the plate circuit.

The 5CX1500A, with 1500-watt plate dissipation, runs considerably under its maximum ratings for amateur operation. During cw tests, it easily delivered a kilowatt into a dummy load.

## references

1. William I. Orr, W6SAI, "2-kW PEP Amplifier for 432 MHz," *ham radio*, September, 1968, p. 6.
2. Robert I. Sutherland, W6UOV, "Design Data for a Two-kilowatt Vhf Linear," *ham radio*, March, 1969, p. 6.
3. *The Radio Amateur's Handbook*, ARRL Staff, 1969 Edition.

**ham radio**

# computer processing slow-scan television pictures

A discussion of computer techniques used to enhance the quality of slow-scan television pictures

Theodore J. Cohen, W4UMF, Howard L. Husted, Paul R. Lintz

**Of all computer applications,** the processing of pictorial information is perhaps the most spectacular. Using a high-speed digital computer and its peripheral output equipment (the line printer, contour plotter, cathode-ray tube display, etc.), it is possible, for example, to transform a multitone photograph into a line drawing, and to obtain various perspective views of the object photographed. You might even simulate the pilot's view of an aircraft-carrier's flight deck together with a "movie" showing a landing on the aircraft.[1] Whatever the job, use of the computer to process and display pictoral and graphical information has provided the scientist, engineer, educator, artist—and yes, the radio amateur—with a highly versatile and accurate tool.

Many readers of *ham radio* may already have an appreciation for the data handling capability of the digital computer; however, the details of picture processing probably remain a mystery. For this reason, we will discuss some basic picture processing techniques. Further, we will apply the processing techniques to slow-scan television pictures to demonstrate the effects on a picture

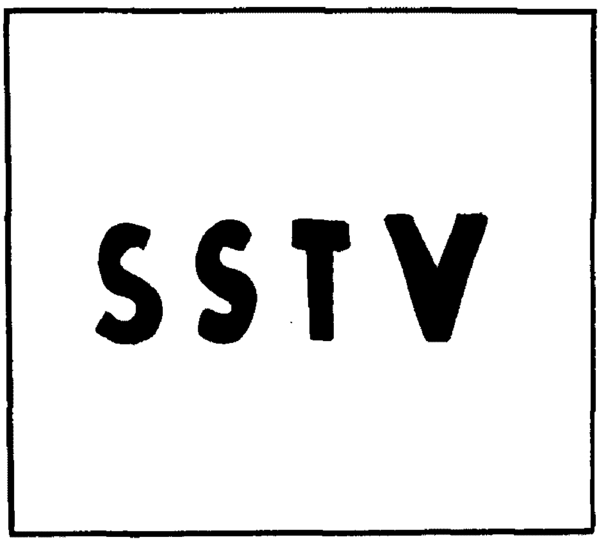



**fig. 1. Sstv test card used by authors to check digitization and computer processing techniques.**

produced by dynamic range compression and contrast enhancement.

## sstv fundamentals

Before examining the various aspects of picture processing, let's briefly review some slow-scan fundamentals. The pictures we processed were produced in accordance with the sstv standard proposed by Macdonald.[2] This standard specifies a slow-scan picture having a line frequency of 15 Hz and a frame period of 8 seconds. Pictures contain 120 lines and consist roughly of 14,400 picture elements. Horizontal sync is transmitted as a 5 millisecond burst of 1500 Hz, while the vertical sync consists of a 30 millisecond burst of 1500 Hz. Video information is transmitted at frequencies between 1500 (black) and 2300 (white) Hz.

## the pictures

The pictures processed came from two sources. The first (fig. 1) was recorded directly from the output of a transistorized slow-scan camera. By using this closed-circuit procedure and a simple four-letter black-on-white card we were assured of having an acceptable picture to test digitization and processing procedures. This test picture was initially recorded on mylar audio tape using a recording speed of 3¾ ips.

The second picture we processed is shown in fig. 2. Originally transmitted on 20-meter ssb by KC4USV, the picture was recorded by W7FEN. The picture's characteristics are less than ideal. Already downgraded by the fact that the multivibrator in the camera at McMurdo was not swinging through the entire 1500 to 2300-Hz video spectrum, the picture was further degraded by atmospheric noise. To the extent that the picture is far from ideal, it presents an excellent opportunity to investigate the capability of our processors.

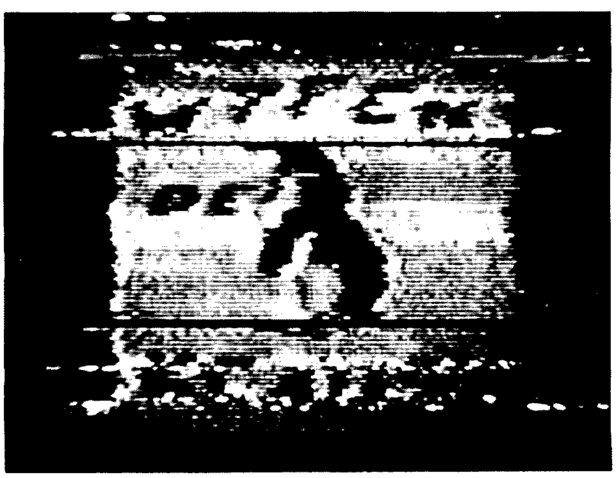


**fig. 2. Single slow-scan frame transmitted on 14 MHz by KC4USV, Antarctica, and received by W7FEN.**

## analog to digital conversion

The pictures were re-recorded on 1-inch, 1.5-mil telemetry tape at a speed of 7½ ips. To satisfy input requirements of the analog-to-digital (A-D) converter, it was necessary to use a playback speed of 0.5 x real time (3¾ ips). Small plug-in RC networks were used to determine the center frequency of the discriminator. At 3¾ ips playback speed the center frequency is 750 Hz; our discriminator's center frequency, however, could not be lowered below 840 Hz.

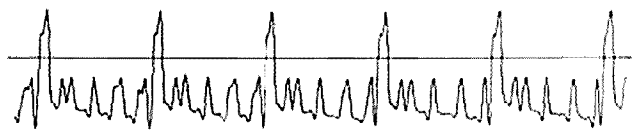


**fig. 3. Successive horizontal lines of the sstv test card.**

The pictures were digitized at an equivalent rate of 2,000 samples per second; this provided a total of 16,000 data points for one complete sstv picture. After A-D conversion, the digitized pictures were placed on magnetic tape.

Oscillograph records of the sstv pictures proved helpful during processing. A sample record showing several lines of the "SSTV" test card is shown in fig. 3. Note that the sync pulses are negative, and that the downward excursions of the positive values display portions of the "S," the next "S," the stem of the "T," and the two sides of the "V." You also can see that each line is brighter on the left side than on the right. This results from the original test card having been illuminated from the left.

## picture preparation

The ideal digitized picture contains two classes of information; sync and video. The viewer is only interested in the video, while the sync insures that each line is displayed properly. The process of preparing this picture for computer processing is one of stripping off the sync and of storing the picture information in some systematic manner. However, problems related to picture digitization and quality must be overcome if we are to effectively process the pictures.

We will restrict the discussion below to the general types of problems involved in preparing a picture for computer processing.

**Identification.** As we digitized many pictures, a label was written ahead of each picture. This label consists of a number and alphanumeric title; the total number of points in the picture, normally 16,000, is also recorded. To find a specific picture the computer searches for a given number and reads the stated number of digital points into memory. The alphanumeric title is also read and is used later to describe the processed picture.

**Picture bias.** Because of our discriminator's center frequency being unavoidably high, the digitized pictures contained a negative offset, confusing the sync and video signals. Thus, a fine adjustment to the dc level was required. This was accomplished by examining the photographic readout of the digitized picture, determining the negative offset, and adding the absolute value of the offset to each picture point.

**Clipping.** We frequently observed large positive spikes in the signal. The spikes use up the dynamic range of the picture, relegating the video information to the obscure background. To meet this problem, the white level of the bias-corrected picture was determined, and all information (spikes) above this level clipped of.

**Sync stripping.** Sync pulses should be distinguished by their negative values and their length in time. At 2,000 samples per second, a 5 msec horizontal sync pulse should contain 10 digital points, while the vertical pulse, being 30 msec long, contains 60 points. Unfortunately, the sync is not stripped easily. Since the sampling rate of 2,000 samples per second is only a nominal value, sync pulses generally contain more or less points than the theoretical 10 and 60; further, we occasionally found positive spikes within the negative sync pulses. Thus, we required that once a negative point was found, the computer interpret this as the first point of a sync pulse. The computer was then to look for the first positive point to appear next (the first picture point of the line), *but* to ignore all positive points which appeared within 8 points of the first negative point.

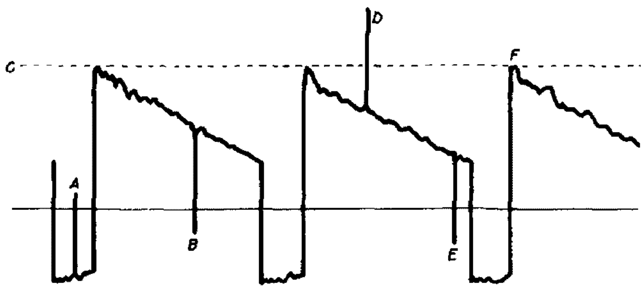



**fig. 4. Hypothetical display of a "worst-case" sstv signal. Fortunately, most of the picture information we processed was free from interference!**

Similar to this problem was the presence of negative spikes within the video portion of the picture. To insure that these spikes did not trick the computer into thinking it had found a sync pulse, we required that once a line of video was found, the computer was to process at least the next 115 points as video.

**Fig. 4** shows the type of digitized information encountered and the manner in which this information was processed.

Note that if the timing of the original sstv signal were perfect, and if the digitization rate were exactly 2,000 cps, the raw picture would consist of 120 lines, each line containing 123 video points. The lines we will process, however, are

made to consist of 120 video points. This is done purposely so that we can display the picture on the computer's line printer in the same manner that one types out a picture using X's, 5's, etc. (quite a popular pastime with the RTTY crowd).

## picture processing

Once the sync has been stripped from the signal and each line filled to 120 points we are ready to begin processing the picture matrix. A matrix is nothing more than an orderly array of numbers; here we allow for a matrix with up to 140 rows (sstv lines) and 120 columns (120 horizontal picture points). Let us perform the following operations on this picture matrix:

1. Quantize the picture elements and plot the picture on the line printer using various symbols (X's, 5's, etc.).
2. Contour the picture (line extraction).
3. Compress the dynamic range and enhance the contrast.

Certainly, computer processing has progressed to the point where the first two operations are easily executed. The third type of processing is a bit more sophisticated, however, and generally requires a computer with a large memory. The computer we used has a rather limited core (32,000 locations), of which up to 16,800 locations can contain picture points. For this reason only a very basic type of dynamic range compression and contrast enhancement will be demonstrated.

## quantizing to n levels

Having formed a picture matrix, one of the simplest forms of picture processing we can perform is to quantize the picture elements and plot the levels using the line printer's symbols. To do this we first determine the maximum value within the matrix and divide every number by this maximum value. Assuming we want N levels in the plot, each point is multiplied by (N-1) and 0.5 added to the product. These numbers are then truncated (numbers after the decimal point thrown away) and a specific symbol assigned to each integer.

As an example of this procedure, consider the following foreshortened sstv line:

2.50 1.25 2.50 5.00 5.00 2.50 2.50 5.00 5.00 2.50 1.25 2.50

The highest number is 5.00; when we divide each number by this value, the line values become:

0.50 0.25 0.50 1.00 1.00 0.50 0.50 1.00 1.00 0.50 0.25 0.50

Let's assume we want to plot the picture on two gray-scale levels, black and white. N is 2, and (N-1) is just 1; thus, multiplying each point by 1 and adding 0.5, we obtain:

1.00 0.75 1.00 1.50 1.50 1.00 1.00 1.50 1.50 1.00 0.75 1.00;

which, when truncated, yields:

1 0 1 1 1 1 1 1 1 1 0 1

If we assign the symbol X when the picture element is 1, and a bland when the element has value 0, the line appears as:

```
X   X X X X X X X X   X
```

Fig. 5 shows the test "SSTV" call card displayed in two gray-scale levels. As you would expect, the printed material requires only the black-white display for effective presentation (the dot between the S's is a blemish on the vidicon target). A printer plot of the Antarctic picture for two gray-scale levels is shown in fig. 6. Despite the fact that the original material was black and white, the two-level display is far from satisfactory because of tonal gradation introduced by transmission. This picture will require further processing.

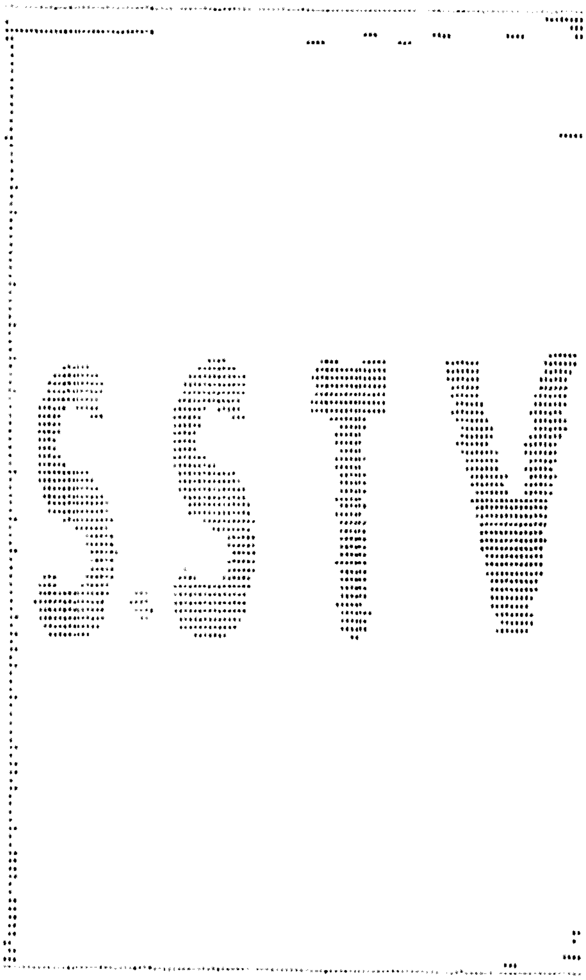



**fig. 5. Line printer display of the sstv test card. Displays of this type are quite popular among RTTY enthusiasts.**

Printer plots of the type discussed above may at first appear to be a novelty; however, they afford the analyst with a good first-order presentation of the picture matrix and allow him to make decisions as to further processing the picture.

Note that the printer-plot pictures are spread in the vertical direction. This is a consequence of the intersymbol spacing being less than the line spacing. However, the advantage of using this form of display for a quick first look far outweighs the geometrical distortion.

One important aspect of the quantizing procedure is that we could now take advantage of the intrinsic redundancy present in the picture. We see from the line of truncated values that data points exist which may be derived from neighboring data points using some simple interpolation procedure. In the line processed previously the fourth through tenth points can each be derived from the third point. All we have to do is transmit data where the values of the picture change, the location of this data, and instruct the processor how to reassemble the picture. Thus, for the line:

1 0 1 1 1 1 1 1 1 1 0 1

we transmit only:

| location | value |
|---|---|
| 1 | 1 |
| 2 | 0 |
| 3 | 1 |
| 11 | 0 |
| 12 | 1 |

and insist that where a value is not specified, the previous value be substituted. Here 10 pieces of data must be transmitted instead of the original 12,

**fig. 6. Line printer display of the KC4USV—W7FEN slow-scan picture.**

and the redundance factor is 12/10 or 1.2. While this is not high, we can easily see that if the redundance factor can be made sufficiently large, elimination of the redundant data points can result in a significant saving in bandwidth.

We can also consider the redundancy between successive pictures as did Ingerson.[3] Here, the entire picture is stored in a computer's memory and only those points which change from picture to picture by more than a minimum specified amount are changed.

## picture contours

Considerable information is contained in a multitoned picture. Yet in many cases we want to retain only the essential features of the picture. For example, it may be our purpose to identify a given pattern within a background of noise; to do this we only need the outline of the pattern. What we want to do is to extract a black-and-white line drawing of the picture—that is, we seek to contour the picture at various amplitude levels (i.e. gray levels). Standard routines are available to contour data, and we will not detail their logic here.


**fig. 7. Computer contour plot of the sstv test card. Note that line extraction causes only the essential features of the picture to be retained.**

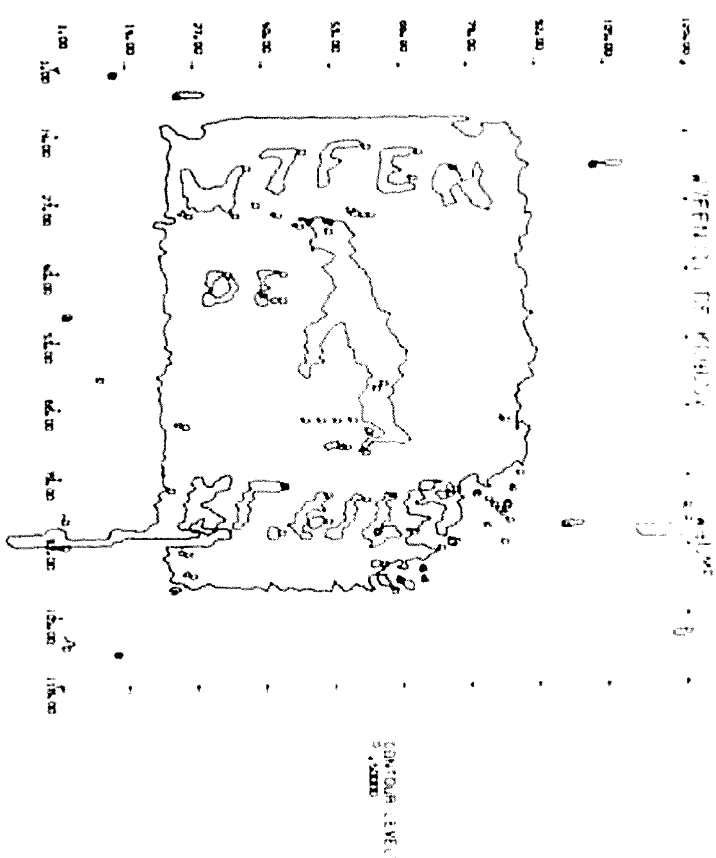


**fig. 8. Computer contour plot of the KC4USV—W7FEN slow-scan frame.**

Contour plots of the "SSTV" card and the KC4USV transmission are shown in fig. 7 and 8. Both pictures were contoured at the 0.5 level (peak level = 1.0). In each case the essential features of the information contained in the original picture have been extracted and outlined.

## dynamic range compression and contrast enhancement

We can consider our picture elements $B_{xy}$ to be formed by the product of an illumination term $I_{xy}$ and a reflectance term $R_{xy}$.[4] The illumination function varies slowly over the picture so it consists of low-frequency components. The reflectance term, however, contains the picture information and is represented by high-frequency components. The brightness at a given point (x,y) in the picture matrix is then the product of the illumination term at (x,y) and the reflectance term at (x,y). That is:

$$B_{xy} = I_{xy} \cdot R_{xy}$$

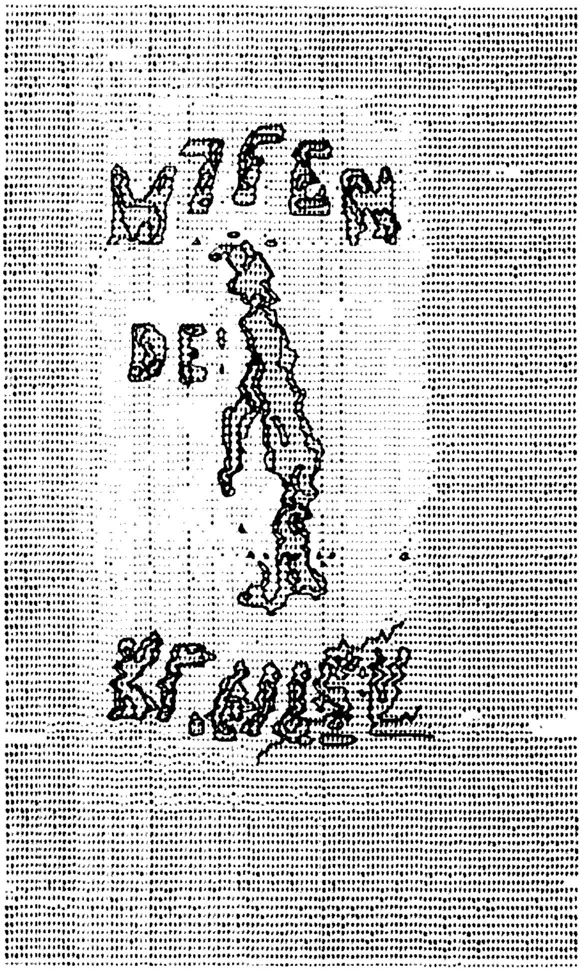


fig. 9. Five-level line printer display of the KC4USV picture. Because of poor contrast the picture can only be defined by using the third gray-scale level (0.625 on a scale of 1.0).

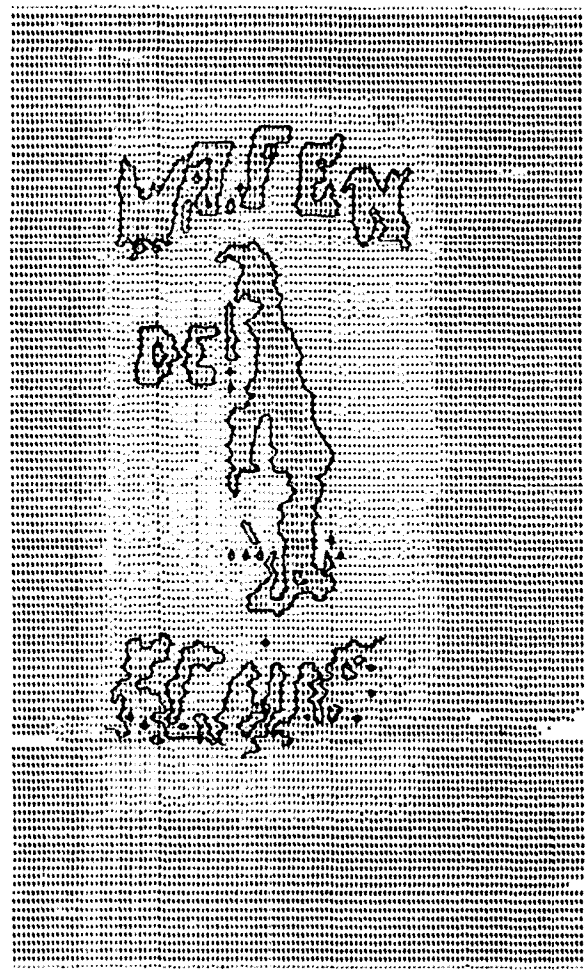
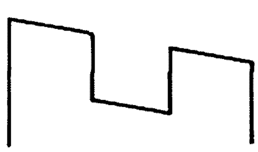
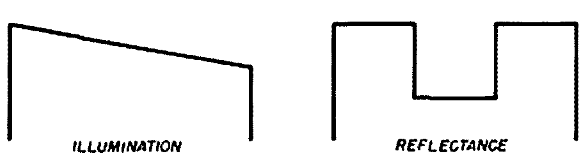


fig. 10. Five-level printer display of the processed KC4USV picture. Because of enhanced contrast an improved picture can now be outlined using only the second gray-scale level (0.375 on a scale of 1.0).

We can easily demonstrate the multiplicative nature of the picture as follows. Consider a picture that consists of a black square on a white background. The picture is illuminated from the left so the right side of the picture is darker than the left. One line from the center of the picture will appear as follows:



But this is just the product of the following two waveforms:




We want to reduce the effect of the illumination term by making it more uniform. This can be done by raising the illumination term to a power $\epsilon$ which is less than 1. On the other hand, to enhance the contrast we want to raise the reflectance term to a power $a$ greater than 1. Thus, the processed picture is represented as follows:

$$B_{xy} = I_{xy}^{\epsilon} \cdot R_{xy}^{a}$$

We conveniently assume that the illumination term can be represented by a plane.

Let's apply the above picture processing technique to the W7FEN Antarctic picture. Fig. 9 is a line printer output of the unprocessed picture; five gray scales are displayed with the lower three enclosed by contours (the contours represent the .125, .375 and .625 levels of a picture for which 1.0 is the maximum

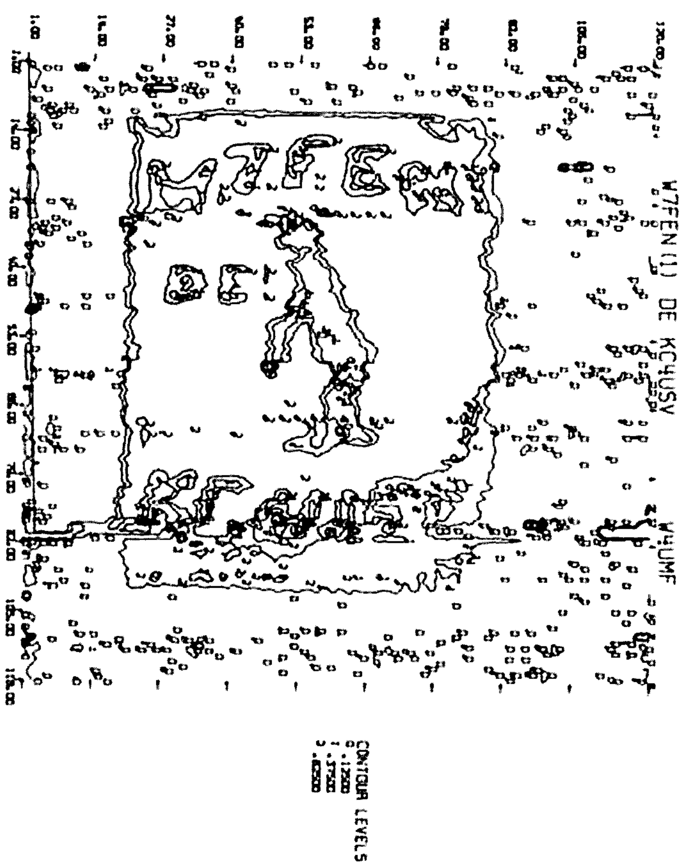



fig. 11. Contour plot of the unprocessed KC4USV picture. Again, it is only after the third gray-scale level is plotted that the picture is defined.

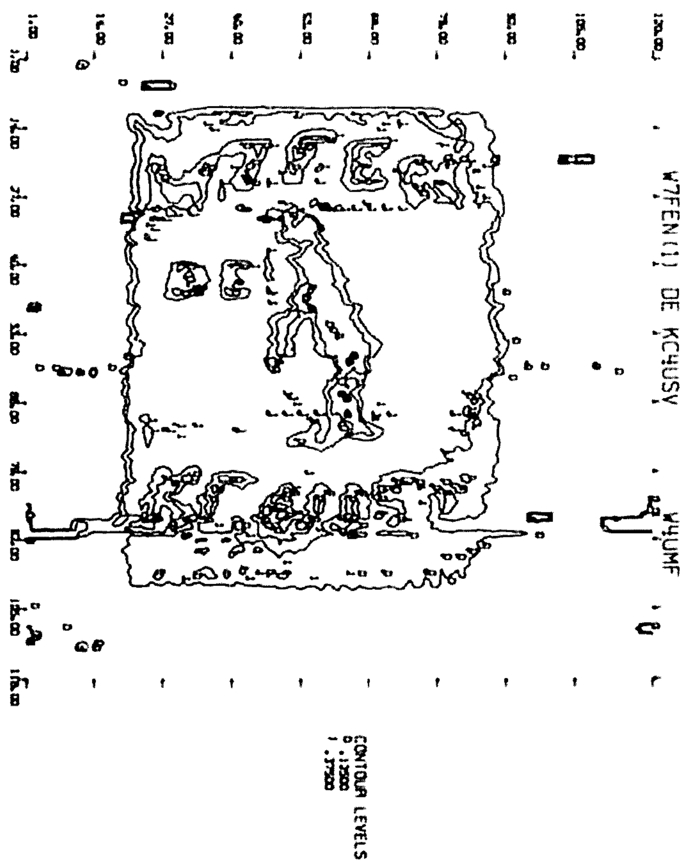



fig. 12. Contour plot of the processed KC4USV picture. As in fig. 10, only the two lower gray-scale levels must be plotted.

picture value). The picture information for the call letters and penguin is very poorly defined. After dynamic range compression and contrast enhancement, however, a more satisfactory picture is obtained (fig. 10). Here only the lower two contour levels (.125 and .375) are needed to define the picture, with the pictorial information now more in relief against a uniform background. A value of 0.5 was used for $\varepsilon$, while the value for $\alpha$ was set to 2.

The differences between the two pictures is portrayed in more detail by the contour plots of fig. 11 and 12. Again, the lower three and two contour levels, respectively, are contoured and plotted. Among other things, note how the penguin is better defined in the processed picture.

## summary

This, then, is a presentation of basic picture processing, one of the most interesting applications of modern computer technology. Using techniques which only a decade ago would have been costly and time consuming to implement the experimenter is now able to see both qualitatively and quantitatively the products of his labor—and he can do so with much saving in time, money and effort. The flexibility of this tool is still being explored with the presentation of moving, colored images but one of many recent advances. Whatever the application, we can be assured that picture and graphical processing by computer will shortly pervade many aspects of our lives.

### references

1. L. E. Harmon, K. C. Knowlton, "Picture Processing by Computer," *Science*, April 4, 1969.
2. C. Macdonald, "S.C.F.M.—An Improved System for Slow-Scan Image Transmission," *QST*, January, February, 1961.
3. T. E. Ingerson, "Applications of Information Theory to Slow-Scan Television," *CQ*, December, 1969.
4. A. V. Oppenheim, "Generalized Linear Filtering," *Digital Processing of Signals*, B. Gold, C. Rader, McGraw-Hill, New York, 1969.

ham radio

# new look in teleprinters

Sam Kelly, W6JTT, 12811 Owen Street, Garden Grove, California 92641

Significant advances have been made in teleprinter development within the last few years. Although they won't reach the amateur market for some time, these new machines have some interesting design trends that will appeal to the amateur rtty fraternity.

This article reviews some of the new teleprinters using thermal, piezoelectric impact, and conventional impact printing techniques. All are designed for mobile use. Interestingly enough, all are also designed for use with separate keyboards. This is because of varied installation and end-use requirements—for example, police-car printers don't require keyboards. Another design trend is toward half-page printers to save space in crowded aircraft cockpits and patrol cars.

Solid-state logic and integral clocks permit a wide variety of input codes and printing speeds. The most commonly used codes are Baudot and the American Standard Code for Information Interchange (ASCII). The latter is similar to Baudot but has eight levels, making possible 128 characters.

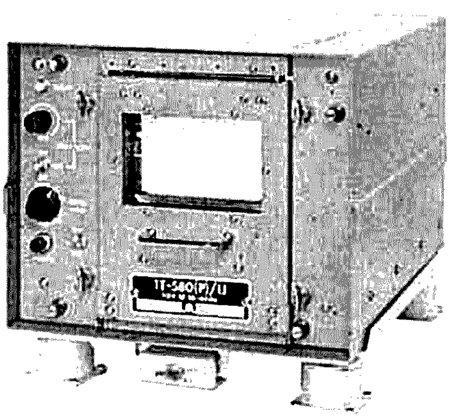
## NCR thermal printers

National Cash Register has developed a technique for true thermal printing. Their printer has no moving parts except in the paper-advance mechanism. The secret to this technique is a special NCR-developed thermochromic paper. The paper has heat-sensitive chemicals dispersed throughout. The chemicals change to a blue-black color when heated by the thin-film print head. The resulting copy will not smear and is insensitive to light. Print quality is as good or better than the best impact printing.

The printer accepts conventional information codes. Speed is in the medium range—up to 300 wpm. The miniature printer has 26 characters per line, while the full-page machine has 80 characters per line.

The incoming signal is processed in a

**Motorola TT-580 miniature teleprinter. The DATA position of the speed switch is for ASCII code; others are for various speeds of Baudot.**



solid-state logic section, which drives the print head. The head consists of a matrix of thermal elements that apply heat instantaneously to discrete dots. On a print command, the desired elements are heated, forming dots on the paper to make the characters. Each character is formed from a 5 x 7 dot matrix. Line feed is automatic upon completion of a line. Since the print head has no moving parts, there's no need for a mechanical carriage return. The printer is extremely quiet, both accoustically and from a radio-frequency interference standpoint.

NCR has also developed a thermal tape printer/reader that handles either thermal or conventional. punched tapes interchangeably.

## piezoelectric printers

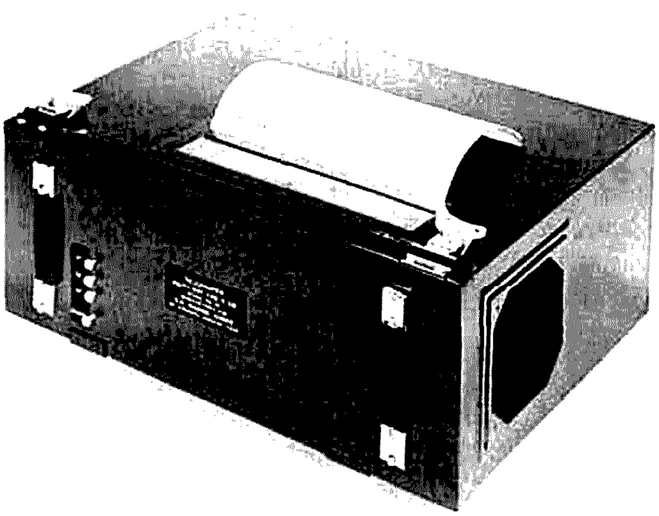
Motorola's TT-580 is classed as an impact printer, but its design is different from conventional impact units. As with all new printer designs, the TT-580's solid-state electronics can be adapted to handle any of the standard input data codes.

The novel feature of the TT-580 is its character-printing method. The hammer consists of a brass plate sandwiched between two 0.009-inch-thick plates of lead titanate zirconate—a piezoelectric device. A voltage applied to the hammer causes it to flex. The hammer flexes in one direction then reverses, resulting in amplitude A (fig. 1).

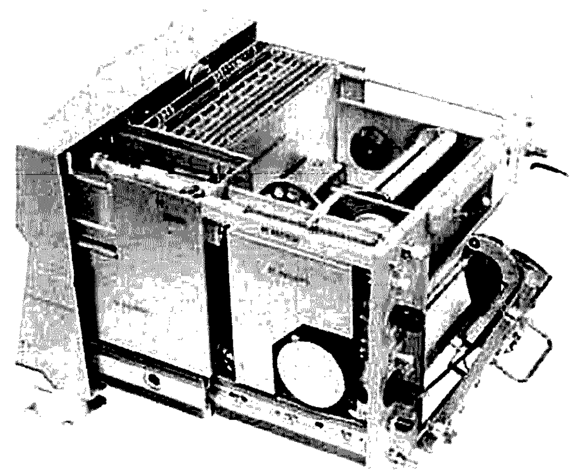
Dots are produced when the hammer impacts pressure-sensitive paper against a helical platen. Characters are formed on a dot-by-dot basis, using a conventional 5 x 7 matrix. The unit uses only six crystals to print all 30 characters. The paper moves in a continuous vertical flow. The printer weighs only 15 pounds; power consumption is a mere 5 watts.

Motorola also makes the VP-100 printer, which is just slightly larger than a cigar box. A special code is used that requires a wider bandwidth than Baudot. However, the code provides communication security and higher immunity to man-made interference. A selective calling feature allows messages to be sent either to individual stations or to all stations simultaneously.

**The Codamite TT-587 printer used by the U.S. Marine Corps.**



**The TT-580 with cover removed.**



## new impact printers

Anyone familiar with World War II machines would be amazed at the clean, simple design of the Codamite printers developed by Bill Johnson, W6MUR,*

*Amateurs who haunt the 20-meter band will surely remember W6MUR's impeccable cw signal—an inspiration for all who aspired to good radio-telegraphy. editor

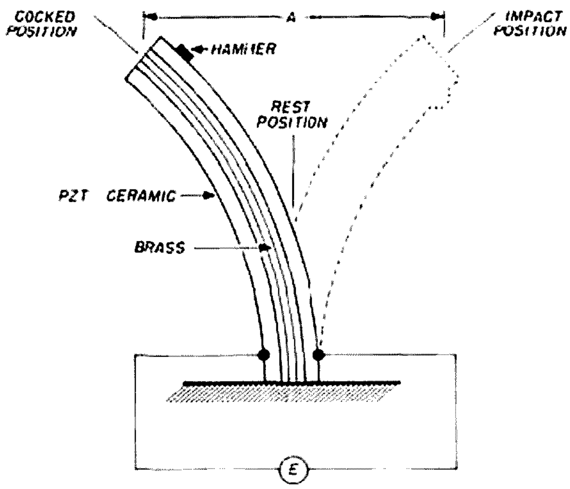



**fig. 1. Piezoelectric hammer used by Motorola. The bimorph bender unit is made from two pieces of PZT ceramic bonded to a brass plate with conductive epoxy adhesive.**

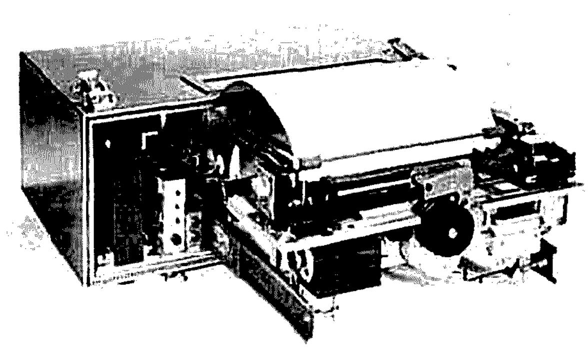
and his staff. These are half- and full-page units for military and law-enforcement use. Their design is a radical departure from conventional impact printers.

The Codamite series uses three precision stepping motors to index the type font, which is in a conventional 8 x 8 octagonal matrix. The type is made from a special metal-plated plastic. The stepper motors are coupled by high-strength, metal-taped pulleys. One motor rotates the type font, a second traverses the printing hammer, and a third translates the font. A miniature solenoid acts as the hammer. The ribbon is a mobius loop, ingeniously re-inked, and very easy to replace.

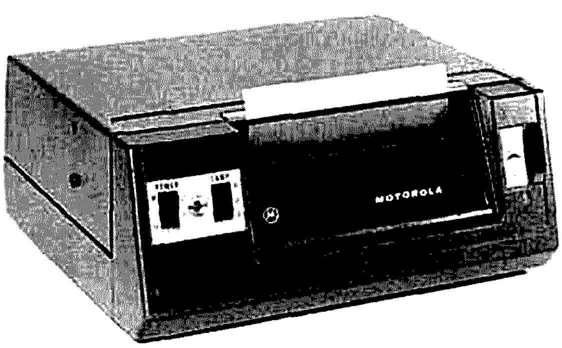
While the new mobile printers are half-page units, carriage return and line feed are designed so that no characters are lost when reproducing full-page copy. The TT-587, shown in the photos, is a special design for tactical military and cryptographic use. Note the features employed to reduce rfi: fingerstock shielding, double enclosure, and screened paper-feed window.

## conclusion

The case for using rtty printers in police cars and other mobile installations is a strong one. Radio frequencies are used to better advantage, message security is obtained, and hard copy provides higher accuracy and a complete record of the message. It's a good bet that printers will be used in most of the country's public-service vehicles within the next few years.



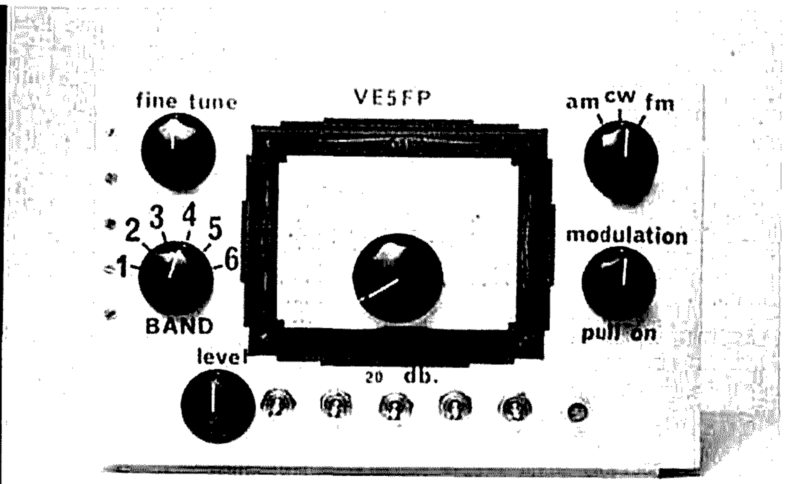
**Inside view of the Codamite TT-587. Unit is packaged to reduce rfi and is ruggedly designed for field use.**




**The Motorola VP-100 miniaturized teleprinter. It uses a special code for communications security and features selective calling.**

The trend in new teleprinters is toward light-weight, compact, easily maintainable designs. The price range of units described in this article is from $800 to over $12,000. However, the day will come when you can have a complete rtty station in the same space occupied by your transceiver and at a comparable price.

ham radio




# a solid-state rf signal generator

This instrument features automatic level control and narrowband fm sweep

J. A. Koehler, VE5FP, 5 Sullivan Street, Saskatoon, Saskatchewan

One of the most useful items of test equipment is a good signal generator. The instrument described in this article has features found in expensive manufactured equipment, yet it can be built in a few evenings for the price of the components.

Solid-state construction results in small size and light weight. The tuning range is from 120 Hz to 95 MHz. An automatic level control and narrowband fm sweep circuit are included. Power supply and modulator are integral in the design.

## description

A block diagram is shown in **fig. 1**. The oscillator is a Hartley in a grounded-collector circuit.[1] The oscillator drives an emitter follower, which provides the low-impedance output. The automatic level control eliminates the need for an output level meter and compensating control, which must be adjusted with frequency change. The output is constant within 1 dB.

Frequency sweep is produced by a varicap across the oscillator tank. Sweep

voltage is generated by a unijunction oscillator, which also provides amplitude modulation. Rf output is varried by a switched attenuator in five 20-dB steps. A manual level control is also provided. Maximum output is 100 mV rms; minimum output is about 0.1 microvolt.

## basic circuit

The oscillator, automatic level control, and attenuator are shown in **fig. 2**. The alc consists of a diode detector followed by a dc amplifier that controls the oscillator's base current. One end of the MV1626 varicap is at dc ground through a 1-meg resistor; sweep voltage is applied at this end. The other end of the varicap receives a variable voltage supplied by the "fine tuning" potentiometer.

The tuning capacitor is a broadcast-receiver variable. If you have one with

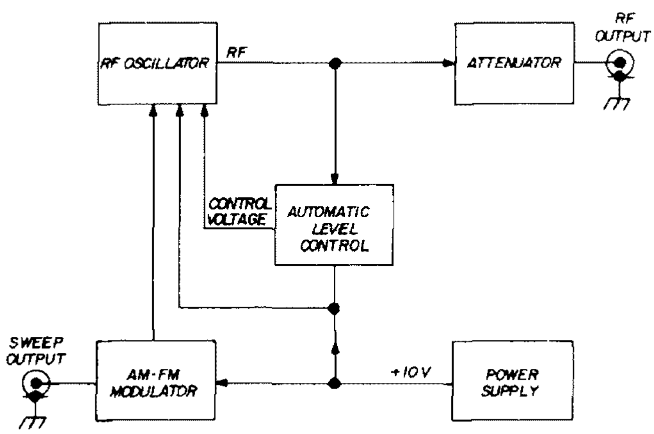



**fig. 1. Block diagram of the solid-state signal generator.**

built-in trimmers, either remove them or set them at minimum capacitance. The tuning dial is a dual-ratio unit made by Jackson Brothers in England. (Miller makes a similar unit in the United States.) To obtain stability and smooth tuning, the same care should be used in construction as when building a vfo.

## construction

Coil and bandswitch layout require some care. For the highest-frequency band, capacitor leads contribute an appreciable amount of the total tank inductance, so they should be as short as possible.

The value of Q1's unbypassed emitter resistor, R1, and L1's tap are determined by trial. Normally, the tap will be between one-fifth and one-quarter of the way down from the top of the coil. The oscillator should just sustain oscillation, and output should be constant across each band (with the alc out of the circuit). These requirements are met by using the optimum combination of L1's tap and R1's resistance. Here's how it's done.

First disable the alc by removing the end of the 18k resistor that goes to Q4's collector. Connect the end of this resistor to +5 volts. Then, for each band, adjust L1's tape and R1's value so the oscillator just sustains oscillation over most of the band. Plenty of overlap will exist between bands, so don't be concerned if the circuit stops oscillating at the low ends. Output level can be checked with a vtvm connected to Q3's gate.

Coil constants are given in table 1. Tap the coil carefully removing insulation from part of one turn, then solder a length of fine wire to it. Take care not to short adjacent turns.

After the oscillator is working well over each band, reconnect the 18k resistor to Q4's collector and adjust the 10k trimpot for about -1 volt at the gate of Q3 (measured with a vtvm). The trimpot will likely have to be touched up a bit to get the same output level over each band.

## the attenuator

A tee rather than a pi attenuator was chosen because it's easier to obtain the required resistance values. The desired values are 40.9 ohms and 10.1 ohms. Standard 5 percent values of 43 ohms and 11 ohms will result in an attenuator impedance of about 55 ohms rather than 50 ohms, but the attenuation will be very close to 20 dB. It's important that the resistors be carbon composition, and the lower the wattage the better. The smaller resistors have a lower inductance, thus

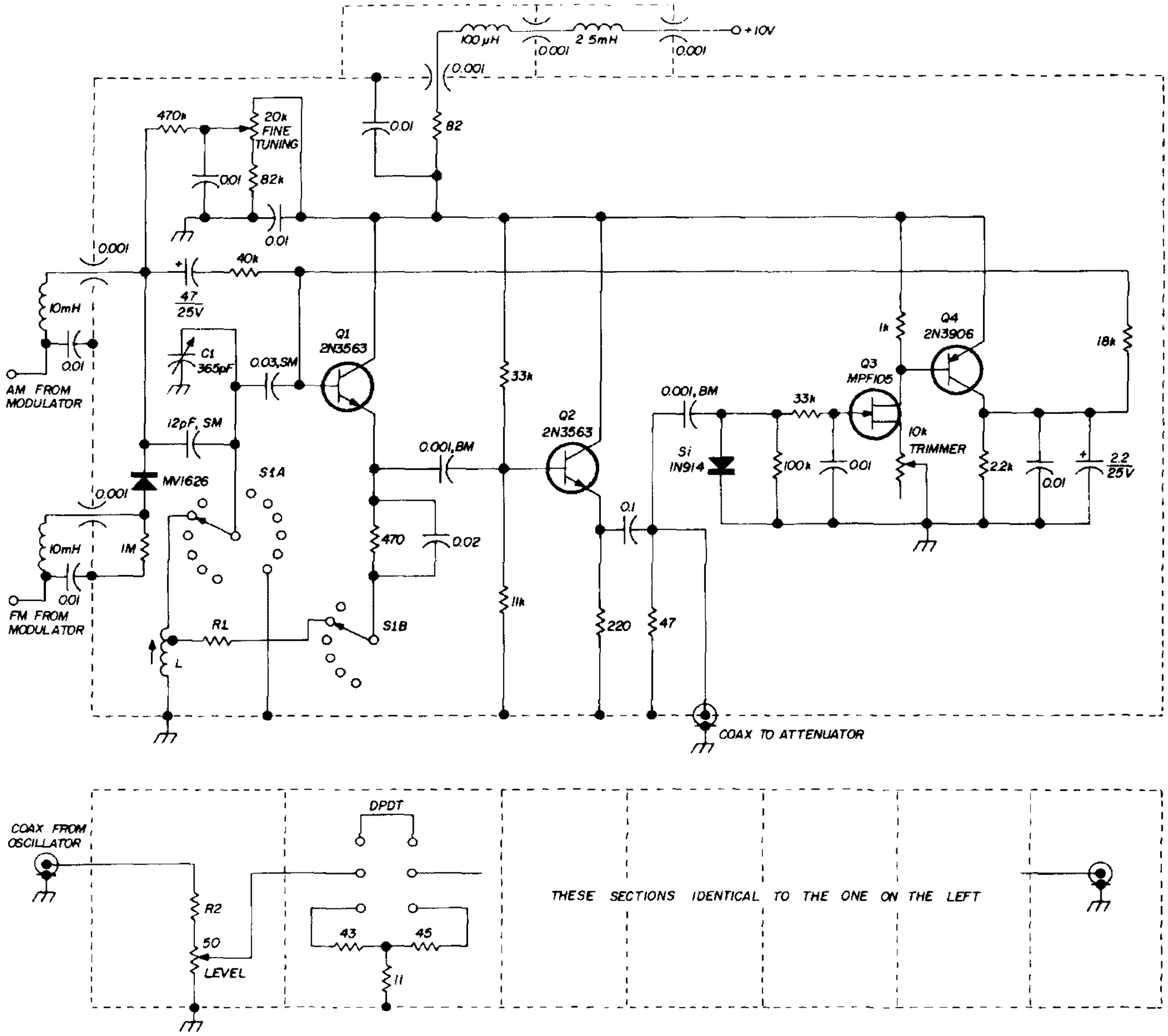



**fig. 2. Oscillator, automatic level control, and attenuator schematic. Oscillator coil tap and R1's value are determined empirically (see text).**

attenuation will be constant at the higher frequencies. I used ¼-watt resistors; the error is small, even at 95 MHz. The level control *must* be a carbon film type.

The value of R2 depends on the rf output voltage produced by the oscillator. Substitute different values of R2 until 100 mV into a 50-ohm load is measured with an rf voltmeter. All attenuation should be switched out, and the level control should be set at maximum. In my circuit, R2 was 120 ohms.

With all five attenuator sections switched in, the output will be 1 microvolt and can be varied to about 0.1 microvolt with the level control. These numbers apply only if the shielding is good. Rf can leak out of very tiny cracks, so the attenuator compartment must be closed very tightly—I soldered mine shut. The coax braid connecting the attenuator to the oscillator compartment should be soldered all around the inside of its entry hole.

Details of the sheet-metal work for the attenuator compartment are shown in **fig. 3**.

## power supply and modulator

The power supply and modulator schematic is shown in **fig. 4**. The power supply uses a voltage doubler circuit with

table 1. Coil data.

| Band | frequency (MHz) | inductance | remarks |
|---|---|---|---|
| 1 | .12 - .28 | 2.5mH | a 4-pie RFC tapped between first two pies |
| 2 | .45 - 1.0 | 350 μH | Miller 9012 |
| 3 | 2.2 - 5.5 | 30 μH | Miller 4407 |
| 4 | 5.5 - 18 | 3 μH | Miller 4404 |
| 5 | 16 - 55 | 0.25 μH | 5 turns on ¼" dia, slug-tuned form |
| 6 | 25 - 95 | 0.025 μH | a hairpin of #14 AWG ½ inch long and ½ inch wide |

a filtered output of 15 volts. The filter drives an emitter follower, whose base voltage is fixed by a zener. Any zener between 10 and 12 volts will be suitable.

The modulator is a simple unijunction relaxation oscillator coupled into a source follower. The source follower output is switched either to the varicap in the fm mode, or through a filter to the base of the rf oscillator in the a-m mode.

For a-m, the modulation frequency is 1000 Hz, which can be set by the trimpot. For fm sweep, the frequency should be quite low—about 5 to 10 Hz.

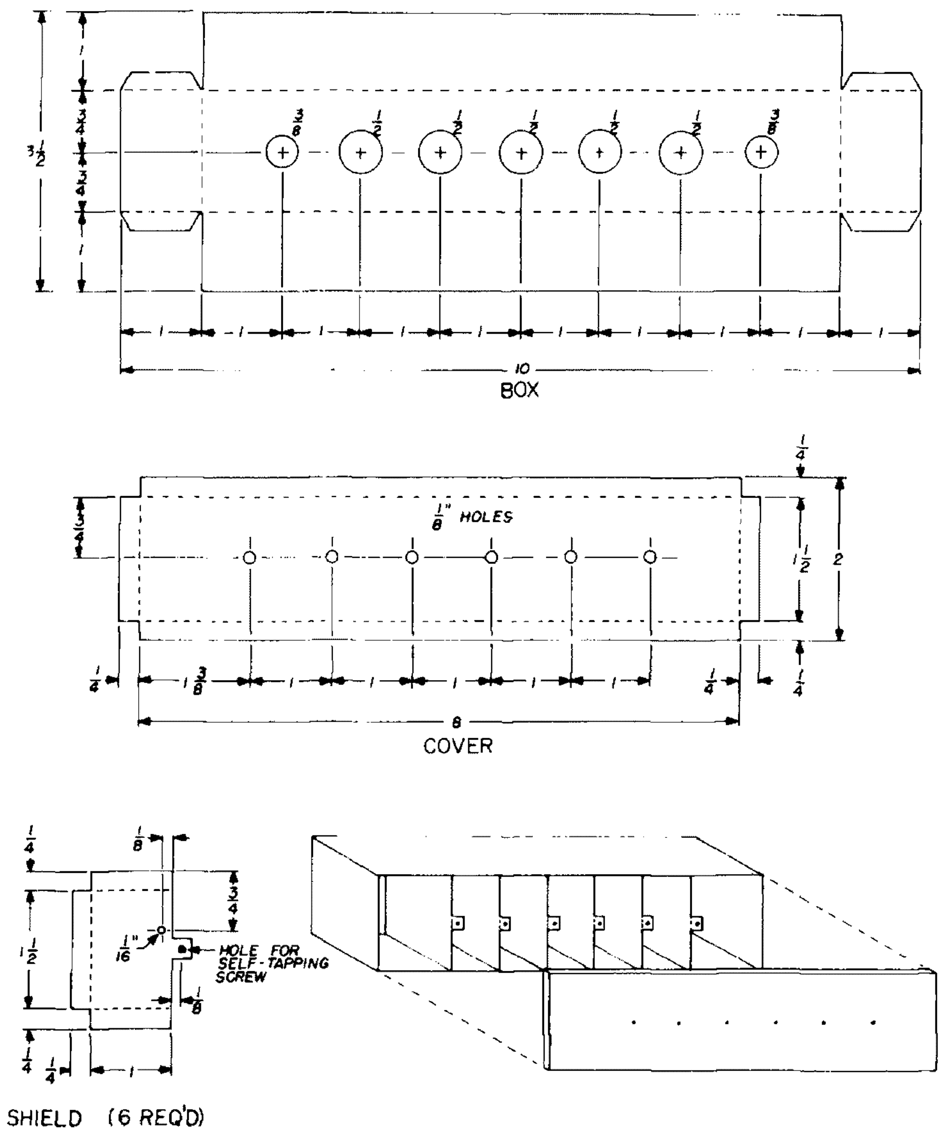



fig. 3. Layout for attenuator enclosure. Material can be thin brass sheet or "tin can" stock.

Because the supply voltage is rather low, you may have some difficulty getting the unijunction oscillator to operate with the required high value of emitter resistance. Use the highest resistor value that sustains oscillation.

The LC filter in the a-m position smooths the sawtooth waveform into a more sinusoidal shape. Ideally, the choke and capacitor should resonate at the modulation frequency. However, it's not too important that the waveform be sinusoidal.

attached to the chassis under the oscillator compartment. The construction of the filter is shown in fig. 5.

The 110-volt ac input enters the chassis through another low-pass filter. I was fortunate enough to find several good line filters at a local surplus emporium. However, a suitable unit could be made by using two more low-pass filters similar to the one in the positive-voltage line. They should be enclosed in a metal box.

The 110-volt ac input should enter the chassis through a male plug mounted

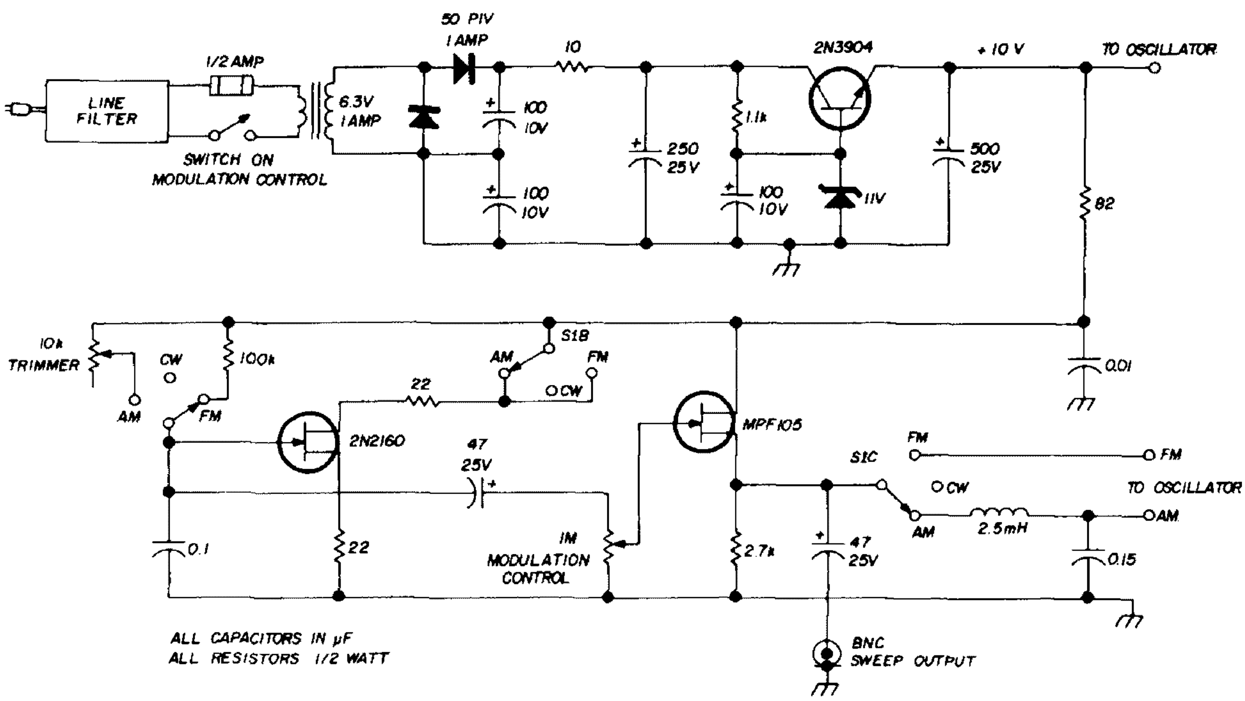



**fig. 4. Power supply—modulator schematic. A simple voltage-doubler, zener regulated supply is used. The modulator supplies either a-m or fm sweep voltage.**

The degree of amplitude modulation and the sweep width are controlled by the "modulation" pot. Sweep voltage is fed to a BNC connector on the chassis rear apron for connection to a scope.

## shielding and layout

The rf oscillator must be well shielded—this means no extra holes in the compartment. In most signal generators, rf leakage occurs primarily through the power-supply circuit into the ac line. To reduce this, the positive supply voltage enters the oscillator compartment through a two-section low-pass filter

inside the filter box. The entire filter assembly should be bolted to the chassis wall.

Make sure the ac line enters the chassis *through* the filter. It's pointless to bring the ac line into the chassis through a grommeted hole and then connect it to the filter. Enough rf will be floating around to couple into any unshielded length of line cord.

General layout isn't critical. The power supply is constructed mostly on a piece of punched board. The modulation-sweep circuit is also constructed on punched board. The power supply is

mounted below chassis; the modulator-sweep circuit is mounted on top of the chassis behind the "modulation" control next to the oscillator enclosure.

Coils are mounted on a vertical plate near the end of the bandswitch. The coils

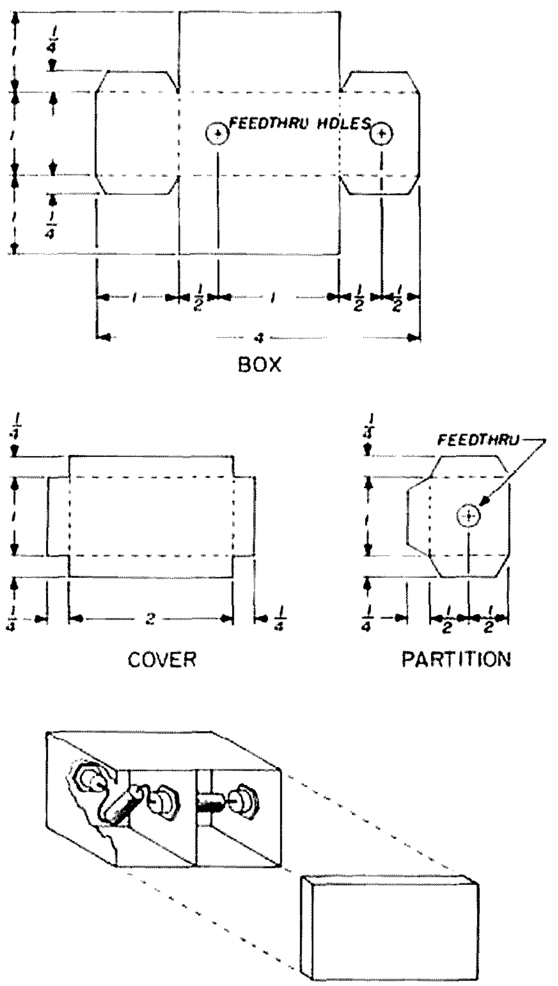



fig. 5. Enclosure details for the low-pass filter. Shielding is important, since the positive supply voltage enters the oscillator through this unit.

extend around and alongside the bandswitch. The emitter follower and alc were built on a small piece of punched board.

The enclosures are made from fairly thin brass. Most of the rigidity is obtained by using a 1/8-inch-thick front panel. A bottom plate is necessary if you don't use a cabinet.

reference

1. L. Nelson-Jones, "Wide-range General Purpose Signal Generator," *Wireless World*, April, 1968.

ham radio

# temperature alarms

## for high-power amplifiers

Some simple circuits to protect expensive components from overtemperature damage

John J. Schultz, W2EEY, 40 Rossie Street, Mystic, Connecticut 06355

**The most expensive component** in a high-power linear amplifier is usually the tube, which can cost from $30 to $60 or more; yet the important precaution of protecting it from overtemperature damage is often neglected.

Although current overload can ruin a tube, the ultimate cause of its destruction is often excessive heat. A simple and dependable temperature alarm is therefore a worthwhile backup to the forced-air cooling system and protective circuits used in most high-power equipment.

This article describes several temperature alarm circuits suitable for linear amplifiers. They were tested, although not all were specifically developed, for such an application. Similar circuits can be used in other equipment where protection against overtemperature is desired.

### temperature sensors

A temperature alarm consists of a sensing device and a control circuit. For low-temperature applications, such as in the common fire alarms, many simple devices may be used. An example is an extremely thin wire that melts when exposed to prolonged overtemperature.

With high-power amplifier tubes, bulb temperature of the order of 200° C are common. A temperature alarm for this application should react to a relatively small temperature rise rather than a drastic increase. Also the sensor must be suitable for use with a specific type of thermistor.

### thermistors

Thermistors are semiconductors whose resistance to current flow is inversely proportional to temperature. They're available in many forms and with many operating characteristics. For high-power-tube protection circuits, the bead-type thermistor is most suitable. The glass bead construction permits operation to around 600° F, which is far higher than would be encountered in a protection circuit for a high-power amplifier tube.

The bead thermistor has the fastest response of all types. Depending on individual specifications, resistance changes from maximum to minimum in about two seconds in response to a sudden temperature change. The small size of the bead thermistor allows placement close to the device to be protected.

The circuits in **figs. 1** and **2** have different features. All can be powered by the ac heater supply in a linear amplifier, so a separate power source isn't required. When the circuits aren't triggered, current drain is only a few milliamperes.

### simple circuits

The circuit of **fig. 1A** can operate a relay to activate a warning light or buz-

zer. (The relay contacts aren't shown in the schematic.) The 500k pot is set so that when the resistance of the nominal 100k thermistor decreases sufficiently, the scr conducts and energizes the relay. No turnoff or reset switch is necessary for the scr, since current through the scr will fall below its holding value. When the thermistor returns to its normal operating temperature, the alarm will turn off automatically. The zener stabilizes the *alarm circuit's operating point against* changes in supply regulation.

## direct audible warning

In the circuit of **fig. 2A**, a dc voltage doubler provides operating power, and a zener is used as a voltage regulator. When the thermistor resistance decreases, the scr conducts, and the 1 $\mu$F capacitor discharges through the scr and the speaker. This causes the scr to return to a nonconducting state. The capacitor then discharges through the 87k resistor. As long as the thermistor resistance remains low enough, the cycle repeats at a rate determined by the RC time constant.

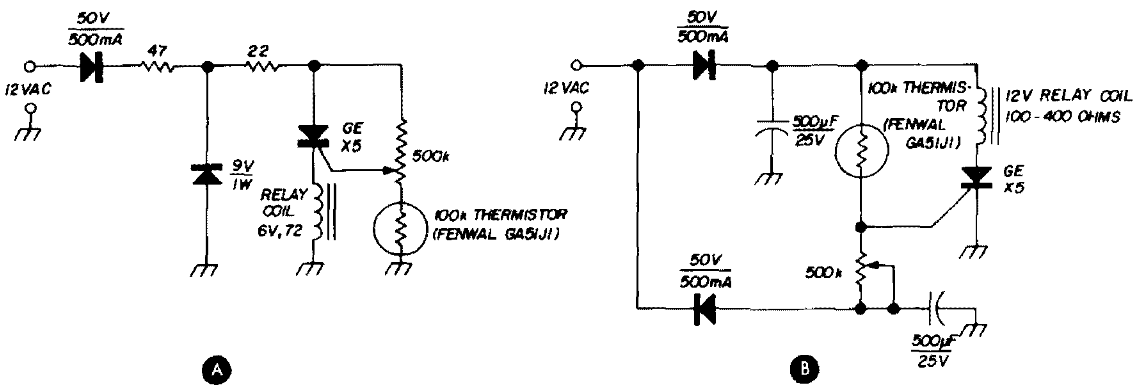



**fig. 1. Two simple temperature alarm circuits that combine SCRs and thermistors. The thermistor is a Fenwal GA51J1; the scr is a 50 volt, 1.6 amp unit from Allied Radio.***

The circuit of **fig. 1B** is extremely simple and stable. No regulator is required, since any changes in source voltage are reflected equally across the series combination of the nominal 100k thermistor and the 500k pot. Again, when thermistor resistance decreases, the scr conducts and energizes the relay coil. With the larger 500 $\mu$F capacitors shown, the alarm circuit will remain on once triggered. It can be turned off (when the thermistor *temperature returns to nor*mal) only by switching off the supply voltage or by using a normally closed pushbutton switch in series with the relay coil.

*Allied Radio Corporation, 100 N. Western Avenue, Chicago, Illinois 60680. The 50 V, 1.6 A SCR is catalog no. 49C3GEX5GE; Fenwal GA51J1 thermistor is no. 60F9909; Fenwal GB32J2 thermistor is no. 60F991.

## large ambient variations

The circuit of **fig. 2B** is useful for large changes in ambient temperature. The simpler alarm circuits, of course, can't discriminate between a rise in bulb temperature and a large increase in ambient temperature. A bridge circuit is used, in which a reference thermistor is located away from the immediate heat area of the device being protected, and a sensing thermistor is placed next to the device.

*An increase in ambient temperature* affects both thermistors equally, and the bridge remains balanced. When the sensing thermistor's resistance falls to a preset value below that of the reference unit, the bridge becomes unbalanced, the scr conducts and the relay energizes. One 5k pot is used to balance the bridge for ambient conditions, and the other is used to set the point at which the alarm is to be activated.

## construction notes

The circuit should be breadboarded first, because construction will depend on the relay and thermistor used as well as conditions for mounting the thermistor next to the device to be protected.

A soldering iron tip moved near the thermistor makes a good temperature simulator. Don't let the iron remain in contact with the thermistor, or it will be destroyed.

The breadboard should be set up in a clear area. Handling a glass-bead thermistor is like working with a dust speck with two leads attached. If the thermistor is lost on a cluttered workbench, forget it!

The sensing thermistor can be located next to a tube, for instance, and the other components can be remotely located. Or the thermistor and alarm-circuit components can be grouped together. Since only dc is involved, there's nothing critical about wiring.

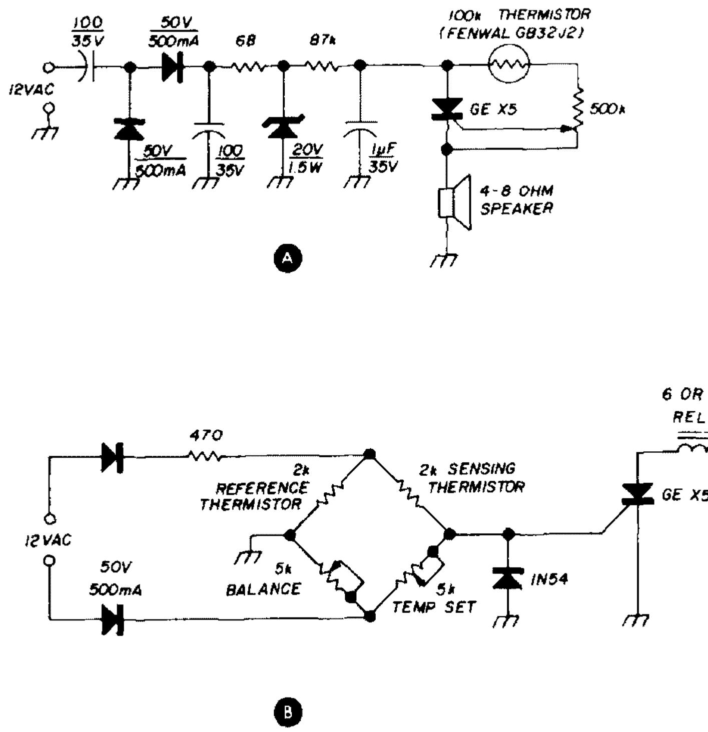



**fig. 2. Two useful circuit variations feature an audio alarm (A), and a comparator circuit (B).**

## thermistor placement

The bead thermistor must be as close as possible to the device to be protected. For a tube, it should be in contact with the seal, because the tube's real temperature rating is slightly lower than that of the glass envelope.

Such placement is difficult and may involve contact with high voltages, so the thermistor can be anywhere near the upper portion of the tube envelope. If the amplifier is subject to movement, as in a mobile installation, the bead thermistor should be held with epoxy cement or a ceramic adhesive.

A similar mounting method can be used with transistors. Special care must be taken to keep the fine wires of the thermistor from contact with the transistor's metal case. Only the glass bead should contact the transistor. A high-temperature cement should be used to place the thermistor and its leads around the area where the bead contacts a metal case.

## operation

With the device to be protected operating normally, set the alarm circuit so it just ceases to operate. If you can safely simulate an overload condition, the alarm can be set to operate for any desired overload.

**ham radio**

# scr regulated power supplies

W.E. Chapple, W4GQC, 309 Bobwhite Drive, S.W., Decatur, Alabama 35601

Silicon-controlled rectifiers as power-supply regulators have many advantages over transistors or vacuum tubes. The scr is an electronic switch that may be either *on* or *off*. Transistors or tubes also may be operated in either of these states, but generally they're used as variable resistors that change value at a rapid rate to maintain constant power-supply output. Such operation causes heat problems that limit maximum output.

Because of low voltage ratings of transistors and low current ratings of vacuum tubes, transistors have been used in low-voltage, high-current regulators and vacuum tubes in high-voltage, low-current regulators. Silicon-controlled rectifiers are now available with ratings of a thousand volts and several hundred amperes. Therefore, scr's can be used in regulated power supplies instead of transistors and vacuum tubes.

Series regulators are more efficient than shunt types, so the following discussion is confined to series-type scr regulator applications.

## operation

The basic scr regulator circuit is shown in fig. 1. It consists of two resistors, an electrolytic capacitor, and an scr. Note that the regulator is inserted between a full-wave unfiltered rectifier circuit and the load. This circuit provides a regulation against changes in load current and a variable output, which is controlled by R2.

The circuit operates as follows. When power is applied, pulsating dc appears on the scr anode and across voltage divider R1, R2. Adjusting R2 provides a voltage for the scr gate.

For convenience, let Ea represent the maximum anode voltage and Eg the maximum gate voltage, as shown in fig. 2. For an scr to conduct, both gate and anode must be positive with respect to the cathode. To turn off an scr, the anode must be negative with respect to the cathode. On the first cycle conditions for conduction are met, and the scr conducts, which charges the capacitor. Near the end

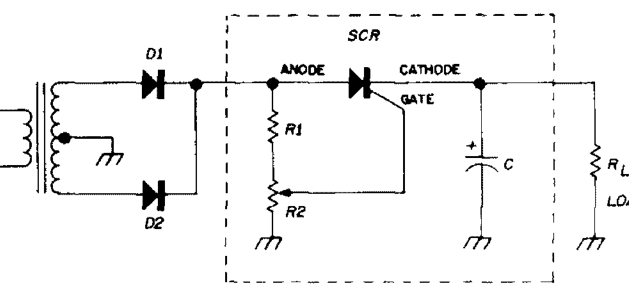



**fig. 1. Basic scr regulator (dashed box). Circuit provides regulated load current and variable output voltage, adjusted by R2.**

of the first cycle, the scr turns off after the anode goes negative with respect to the cathode. The capacitor voltage keeps the cathode at a positive potential with respect to ground. If the capacitor voltage is larger than Eg, the scr will not conduct on the next cycle: in fact, several cycles may pass before the voltage on the capacitor drops below Eg. When this happens, the scr again charges the capacitor, and the process repeats.

The output is limited approximately to Eg, with a sawtooth ripple component determined by R1 and C.

Increasing C increases filter action by increasing the time constant, which prevents the capacitor from charging to a voltage much higher than Eg. Output voltage can be either increased or decreased by adjusting R2. R1 prevents

excessive gate current when R2 is adjusted for maximum output. Changes in RL will not vary output voltage. Higher load currents increase sawtooth ripple frequency to a maximum of 120 Hz for 60-Hz power. Minimum ripple frequency can be smaller than 1 Hz with large values of C and RL, which means the scr conducts once every second or so. Remember, the sawtooth ripple component is caused by the scr conducting.

## low-voltage supplies

Suppose the transformer of **fig. 1** is rated at 100 watts, and Ea is 30 volts. If R2 is adjusted for 20 volts, the load could draw 5 amperes. This, of course, assumes the transformer is 100 percent efficient and no power is lost in the scr. If output is reduced to 10 volts, the load could safely draw 10 amperes. The load power is independent of output-voltage amplitude. For high load currents, the scr must be mounted to a heat sink which, in most cases, could be the chassis. The scr must be electrically insulated from the heat sink with mica washers. The output can be regulated for changes in line voltage with a zener across R2. If you want more than one output voltage, another regulator can be connected to the junction of D1 and D2 in **fig. 1**. The potentiometer could be adjusted to give the second output voltage. Since output is a function of the potentiometer setting, R2 can be calibrated in volts, eliminating an output voltmeter. R1 and R2 aren't critical and can be determined from the minimum and maximum scr gate currents. As previously stated, R1 limits the maximum gate current if R2 is adjusted for this voltage. The capacitor should be as large as practical.

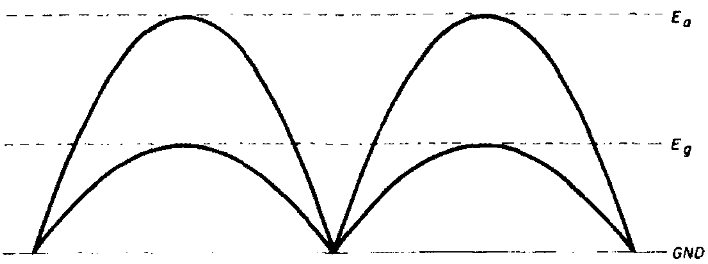


**fig. 2. Output waveforms demonstrating scr operation.**

## high-voltage supplies

The techniques discussed for low-voltage supplies may be used for high-voltage power supplies—with a few precautions.

The gate-to-cathode potential could be exceeded after the capacitor is charged. For instance, assume the capacitor is charged to 200 volts. When the ripple voltage is zero, the gate will be 200 volts *below* the cathode. This will cause a short to develop at the gate-to-cathode junction. To prevent this, a diode is inserted in series with the gate (**fig. 3**). The scr peak inverse rating must be equal to, or greater than, Ea of **fig. 2**.

If the adjustable output feature isn't needed, and you wish some degree of regulation for load-current changes, the circuit of **fig. 3** provides regulated outputs of 800 and 300 volts. With loads

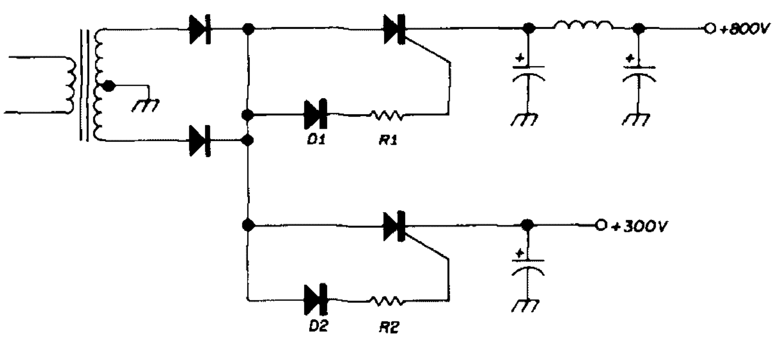


**fig. 3. An scr-regulated supply without adjustable output control. R1, R2 can be determined with a decade resistance box; D1 and D2 prevent gate-to-cathode short circuits.**

removed, the regulators prevent the output voltages from increasing. R1 and R2 are best determined by using a resistance decade box: resistance is changed until the required output voltage is obtained. This allows transformers with higher voltages than required to be used, since output voltages can be reduced to any value by adjusting the gate resistors.

## practical circuit

The schematic of **fig. 4** is a regulated supply that provides useful voltages for many amateur applications. The precautions about mounting the scr's to heat sinks should be observed. The transformers, of course, occupy the most space. They can be mounted on metal chassis, which can also serve as a heat sink for the scr's. The other components can be mounted on a punched board, located vertically to save space. The 5k pot in the low-voltage supply can be calibrated directly in volts, as suggested previously.

Don't forget to insulate the scr's from the metal chassis with mica washers. If a cabinet is used, it should be well ventilated.

## conclusions

These techniques have been used in several power-supply applications. Since scr's are either on or off, it's a simple matter to determine if the regulator is working. If the regulator fails due to the scr shorting, output will rise to the maximum value.

Several output voltages may be obtained by connecting additional regulators to the rectifier outputs. The voltage-control pot may be calibrated directly in volts, eliminating the need for an output voltmeter. Changes in load current don't appreciably change output voltage. A zener across R2 (**fig. 1**) will provide regulation for line voltage variations.

**ham radio**

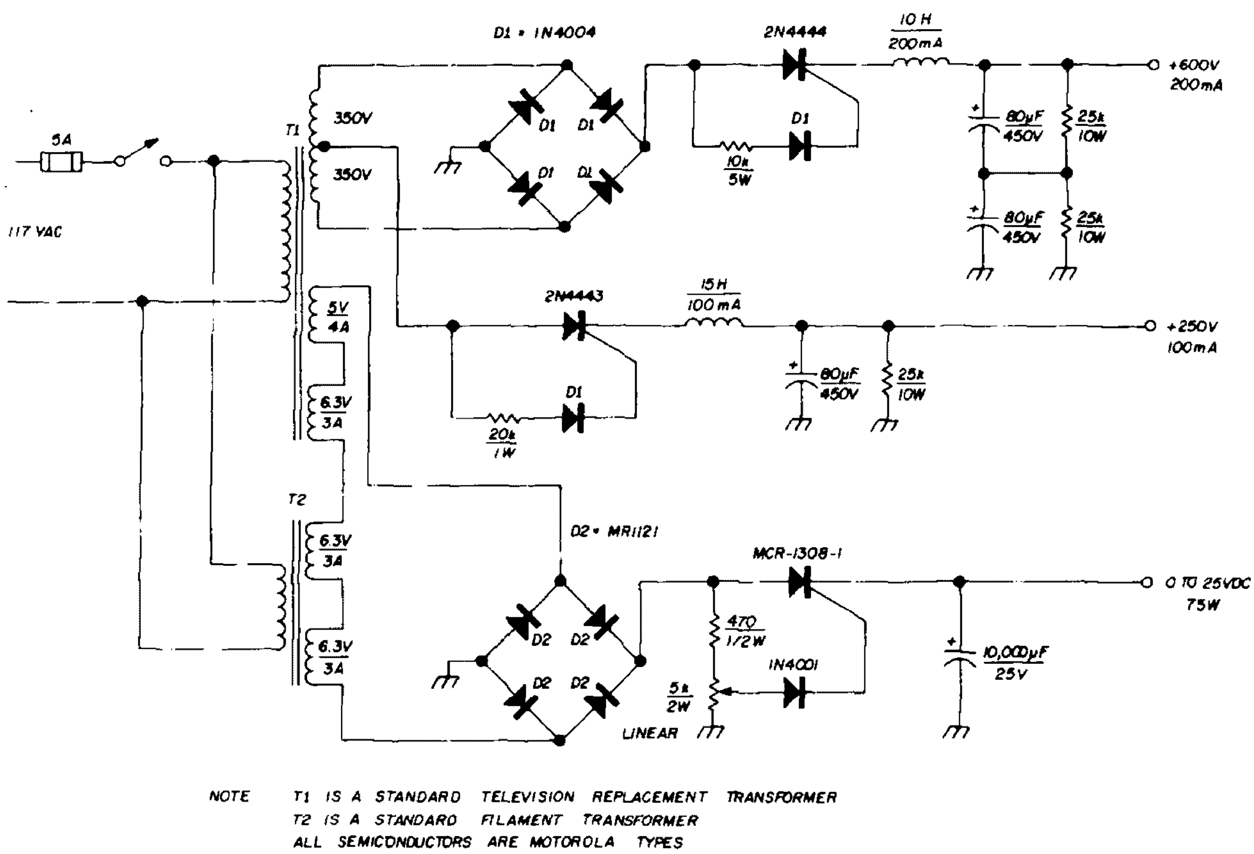



**fig. 4.** Practical power supply using scr's. The pot in the low-voltage unit can be calibrated directly in volts, eliminating an output voltmeter.

# microwave hybrids

## and couplers

### for

## amateur use

Circuits borrowed from the microwave domain have many uses at the lower frequencies— here are some applications

Henry S. Keen, W2CTK, 64 Schuyler Drive, Commack, New York 11725

**Microwave hybrids are extremely** versatile devices. They have many applications not necessarily restricted to the microwave region. This article explains how these circuits may be put to use at the higher amateur frequencies, where communication may be enjoyed without the interference and noise created by thousands of commercial kilowatt transmitters.

When hybrids are mentioned, many hams think of bifilar-wound coils on toroidal forms. However, the circuits described here may be constructed from coaxial line for vhf or uhf use. For the higher frequencies, they may be constructed using stripline, microstrip, or waveguide techniques.

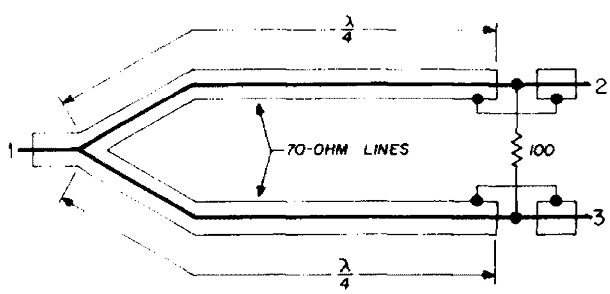
Three devices are considered:

**1.** The half hybrid (**fig. 1**). This is a degenerate form of a 4-port device that may be used as a power combiner or divider.

**2.** The branch directional coupler, **fig. 2**, which is a quadrature hybrid with some interesting applications for moonbounce work and ssb.

**3.** The coaxial rat race (**fig. 3**). Sometimes called a 180° hybrid, this circuit may be used as part of a balanced modulator or to match or balance two equal loads (as in combining equal sections of an antenna array).

*fig. 1.* **The half hybrid, useful as an isolator between two power sources. Impedance at each port is 50 ohms.**



## the half hybrid

This is the simplest of the devices described. It consists of a Y or T junction, two ¼-wave matching transformers, and a bridging resistor.

If the half hybrid is fed at port 1 (fig. 1), the signal will divide equally between ports 2 and 3. Because no phase difference exists at ports 2 and 3 when properly terminated, no voltage appears across the resistor; therefore no power is absorbed. If, however, an imbalance exists due to a mismatch, part of the signal will be absorbed by the resistor and part will be reflected to the generator. If the generator impedance is 50 ohms, it will absorb the reflected portion. The isolation between output ports is independent of the match provided by the loads.

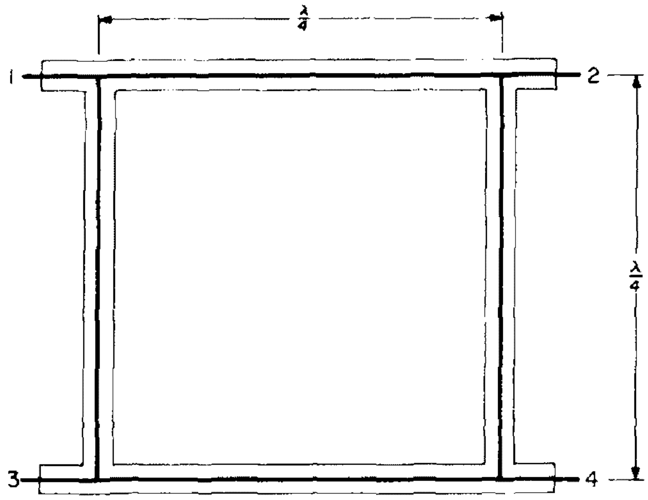



fig. 2. The branch directional coupler. This device divides power between two matched loads.

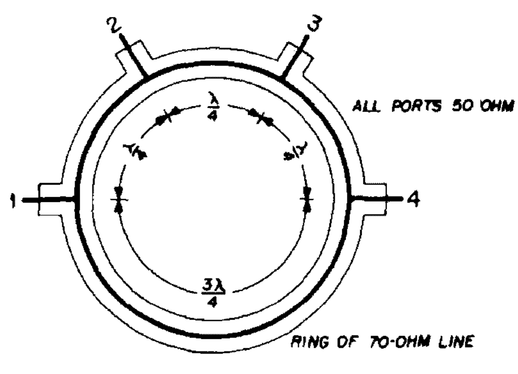



fig. 3. Coaxial rat race, or ring coupler. It can be used to balance similar sections of an antenna or as a balun.

Offhand the source of this isolation may seem vague; but if the circuit is redrawn as in fig. 4, the path from port 2 to port 3 resembles the familiar bridged-T network. In this circuit, a signal at port 2 will be nulled at port 3. Therefore, the

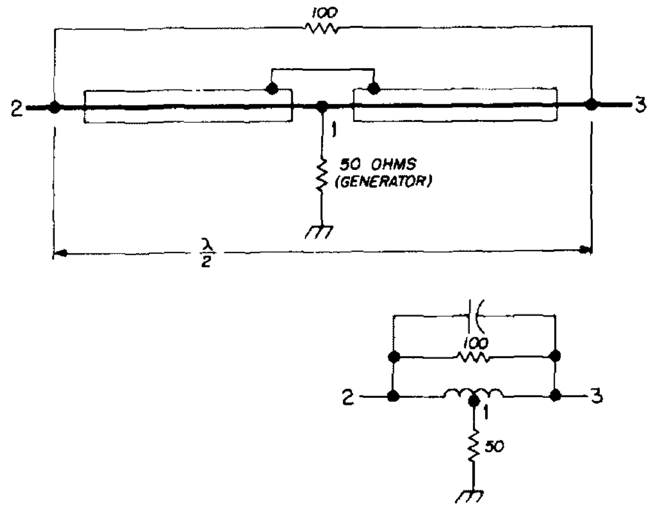



fig. 4. Equivalent circuit of the half hybrid. Note the resemblance to the bridged-T network.

load impedance at each port is not a factor in the isolation between ports.

The half hybrid may be used to provide isolation between two power sources, such as a pair of power transistors. A fail-safe arrangement is thus obtained, whereby failure of either component will not affect the load presented to the other unit. Power output will decrease by 6 dB as input power will dissipate in the bridging resistor, but loading conditions presented to the source will remain unchanged.

In applications requiring high reliability during prolonged unattended operation (as in fm repeaters), half hybrids as combining networks offer a passive means of ensuring uninterrupted service without resorting to complex switching mechanisms.

## branch coupler

The branch coupler is a 4-port device. It divides input power between two matched loads. The isolation between two input ports is a measure of the match provided by the loads. A 90° phase

difference exists between the signals at the two output ports, because one signal travels ¼ wavelength farther than the other. This device can be used for sampling a portion of the signal for reference or comparison purposes.

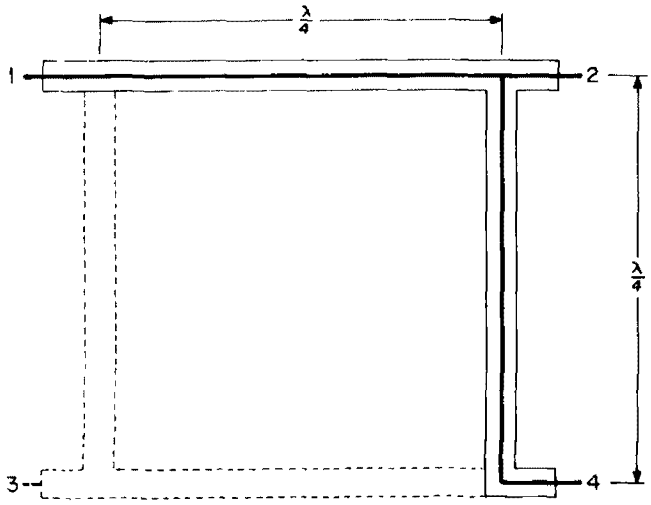



fig 5. The branch coupler. If port 3 is short-circuited, branches 1-3 and 3-4 may be removed.

## analysis

To understand the design principles of the branch coupler, consider the case of the 3-dB version, in which the input power is divided equally between the two load ports.

In a perfectly matched coupler no signal exists at port 3, so this port can be short-circuited without affecting power distribution (**fig. 5**). This would make branches 1-3 and 3-4 shorted ¼-wave stubs, shunted across ports 1-4. Thus they may be removed, leaving only branches 1-2 and 2-4.

If power is to divide equally between ports 2 and 4, port 4 must present a 50-ohm load at port 2. The characteristic impedance of branch 2-4 must therefore be 50 ohms, thus establishing an impedance of 25 ohms at port 2. In order to match this to a 50-ohm input, branch 1-2 must have a characteristic impedance of 35 ohms, which can be obtained with two 70-ohm coaxial line sections in parallel.

When the network is "reassembled," branch 1-3 will be the same as branch 2-4; while branch 3-4 will be the same as branch 1-2. For a general solution of the design, the branch impedances will be:

$$1\text{-}3 = 2\text{=}4 = \sqrt{\frac{\text{Power (2)}}{\text{Power (4)}}}$$

$$1\text{-}2 = 3\text{-}4 = \sqrt{\frac{\text{Power (2)}}{\text{Total power}}}$$

## applications

Several applications of the branch coupler are of interest for amateur work. For example, a 3-dB coupler can be used as a phasing power divider to feed a circularly polarized antenna. Another use would be as a 90° phase shifter for phasing-type ssb generators.

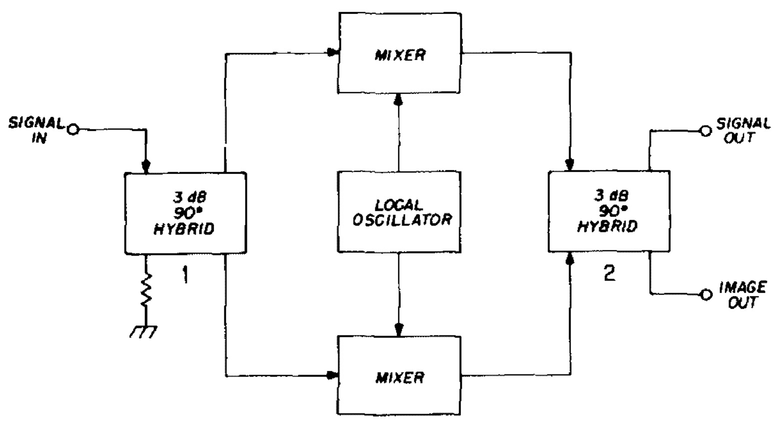



fig. 6. The branch coupler used to phase out front-end image in ssb receiver applications.

Let's first consider the power divider.

If a signal fed to port 1 produces clockwise phase rotation, feeding port 3 will produce counter clockwise rotation. If both ports are fed simultaneously, linear polarization will result. A line stretcher in one of the inputs would permit adjustments to any desired phase angle.

A received signal of the same polarization as that transmitted would appear at the same port from which it was transmitted. A signal of opposite phase rotation, such as a reflected signal, would appear at the other port.

In microwave applications, isolation from 40 to 50 dB has been obtained under ideal conditions. The thought occurs that this idea might be useful for moonbounce work; however, I have no information as to how much the circular polarization would be degraded.

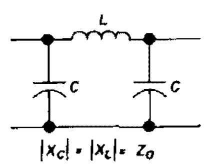



**fig. 7. Synthesized ¼-wave line of impedance Zo.**

## single sideband

Single-sideband phasing techniques have been used in microwave receiver design to phase out the image signal. This method also offers a theoretical 3-dB reduction in front-end noise.

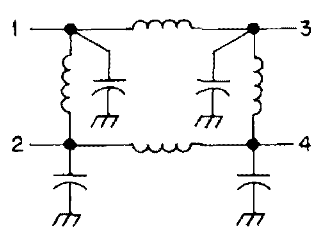



**fig. 8. Lumped-constant equivalent circuit of the branch coupler.**

A block diagram of such a system[1] is shown in **fig. 6**. The second 3-dB hybrid operates as a combining network designed for the intermediate frequency. Balanced mixers could be used to cancel the noise contributed by the local oscillator.

## impedance synthesizer

At lower frequencies, the branch coupler may be synthesized with appropriate values of L and C. An equivalent ¼-wave line may be constructed for any desired characteristic impedance (**fig. 7**). The absolute values of each reactance at the design frequency should equal that of the line being synthesized. The capacitors in the final version, **fig. 8**, are identical in value.

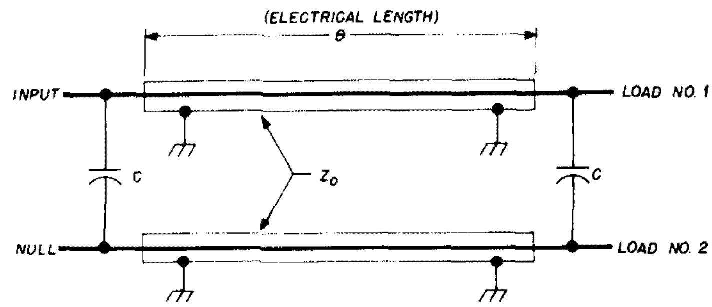



**fig. 9. Capacitively-coupled hybrid is another form of the 90° or quadrature hybrid.**

## coaxial rat race

The standard form of the 50-ohm rat race, or ring coupler, is shown in **fig. 3**. It consists of a closed loop of 70-ohm line with a circumference of three ½ wavelengths. The four ports are located ¼ wavelength apart, with first and fourth ports connected by a ¾-wave line.

A signal fed to port 1 divides in two; each half travels around the loop in opposite directions. The path to port 4 is ½-wavelength longer than that to port 2, so the two signals arrive at their respective loads in phase opposition. Port 3, located midway between the two loads, will therefore receive no signal. The loads must be identical for this cancellation to occur.

As a matter of interest, both loads can be removed, leaving only the loop with ports 1 and 3. Cancellation will occur at the center frequency. This dual-path structure is known as a re-entrant filter.

If the signal is fed to port 3, the two loads will be fed in phase. Any in-phase reflections of equal magnitude from ports 2 and 4 will arrive at port 1 out of phase and will therefore cancel. If the loads are unequal, and the reflected signals differ in amplitude or phase, or both, then cancel-

lation will be incomplete, causing a signal to appear at port 1. In some applications, a matched load may be placed at the odd port to absorb the imbalance.

## applications

The rat race offers an excellent means of adjusting signal balance between similar sections of an antenna array. A detector-indicator, such as a receiver with an S-meter connected at port 3, would show imbalance between array sections. Identical lengths of transmission line must be used between ports 2 and 4 and their respective loads to avoid complications due to phase differences.

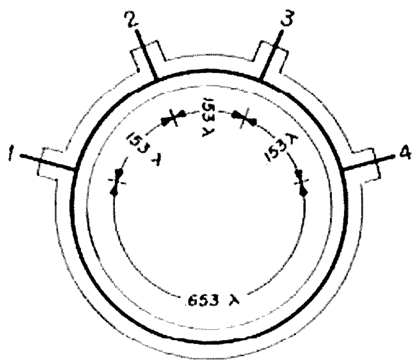



**fig. 10. Coax rat race constructed with 50-ohm line.**

The rat race also functions well as a balun. When used for this application, the balanced load impedance is twice that of the coax input line, and port 3 is usually grounded.

## capacitively-coupled hybrid

The capacitively-coupled hybrid shown in fig. 9 is another form of the 90° or quadrature hybrid. Coaxial line of any convenient characteristic impedance can be used in its construction as long as the correct line length and proper coupling capacitances are used. The electrical length of the line elements is computed from:

$$\text{Coupling (dB)} = -20 \log_{10} \cos \theta$$

The reactance of the coupling capacitors is:

$$X_c = Z_o \tan \theta$$

Thus, for a 3 dB hybrid using common 50-ohm coaxial cable, the lines would have an electrical length of 45°, and the coupling capacitors would each exhibit 50 ohms reactance.

## 50-ohm rat race

If you don't happen to have any 70-ohm line handy, you can make a rat race with 50-ohm line, which is suitable for spot frequency or narrowband work. In this version (fig. 10) ports 1 through 4 are separated by 0.153 wavelength of 50-ohm line. The long side is 0.653 wavelength, taking into account the cable's velocity factor.

At lower frequencies, the rat race is replaced by a center-tapped transformer. In the higher-frequency regions, where waveguide is used, a device known as the "magic Tee" performs the same function.[2]

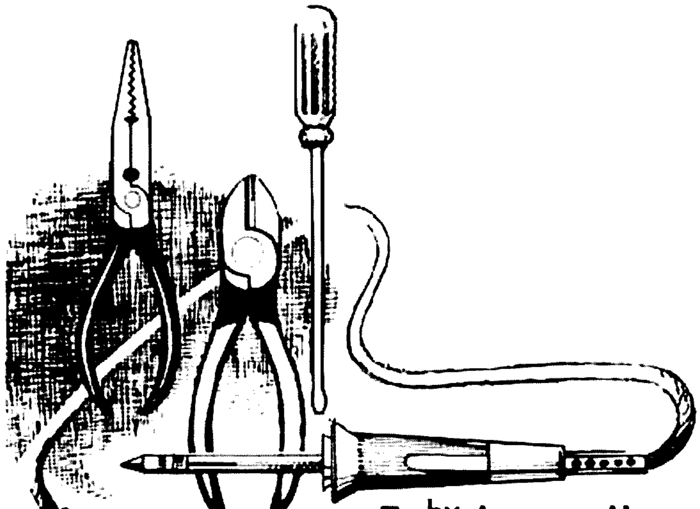
In all regions of the radio spectrum, hybrid devices exist in one form or another, which can contribute much to the versatility of equipment design.


references

1. M. Loss, "Phasing Unwanted Images Out of Microwave Receivers," *Electronics*, July 12, 1965.
2. Keith Henney, "Radio Engineering Handbook," Fifth Edition, McGraw-Hill, New York, p. 6-31.


ham radio


**The DXer's dream.**



by larry allen

# the repair bench

## doing your own transistor tests

To hear some guys tell it, a transistor is the easiest thing in the world to test. But others don't agree. A transistor to them is still a mystery.

Well, the truth is, most transistors can be tested without complicated equipment, gimmicks, calculations, or formulas. To keep it simple, there are just two basic things you need to find out about a transistor: (1) Does it work at all? (2) How well?

### transistor parameters

That word "parameters" scares off a lot of hams. It conjures up complicated graphs with bent lines and long formulas with Greek symbols and big and little letters. All the word actually refers to is *conditions of operation*.

One transistor manual lists 103 possible parameters. They're great for a transistor designer. But a lot fewer is plenty for testing on the repair bench. In fact, I won't even use the term "parameters." Instead I'll just tell you about the voltages, currents, and resistances that tell you how a transistor is doing.

I'll start with the diagram of a simple transistor stage in **fig. 1**. This is a grounded-emitter amplifier—probably the most common transistor stage in use today.

The transistor is npn. Bias is forward when the base is slightly positive with respect to emitter. The collector is "far" positive with respect to emitter.

A pnp transistor takes negative voltage on the base to forward bias the emitter-base junction. That's not necessarily a negative voltage to ground, but to emitter. The collector of a pnp operates "far" negative from the emitter.

### which way is up?

Some hams I've talked to about transistors seem confused by operating voltages. One key to understanding is knowing how to describe the voltages.

For example, in fig. 1 if the base

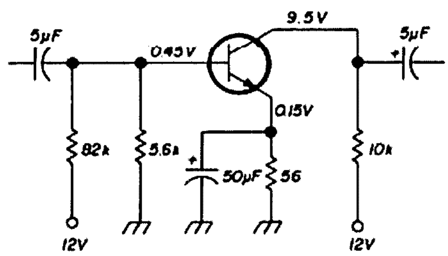



**fig. 1. Common-base transistor amplifier is popular in amateur equipment.**

voltage changes to 0.1 volt, it has obviously become less positive. That means less positive with respect to wherever you're measuring from, and for most measurements that is ground.

Look at the same voltage with respect to the emitter. As it's labeled on the diagram, the base is normally more positive than the emitter by about 0.3 volt. (The emitter is 0.15 volt, and the base is 0.45 volt; between the two is 0.3 volt, the base more positive than the emitter.)

Know what that means. "More negative" is exactly the same thing as "less positive." And "more positive" means the same as "less negative."

If the base voltage in **fig. 1** drops to 0.1 volt, the voltage relationship between base and emitter changes. The difference is then 0.05 volt (0.15 minus 0.1 equals 0.05), but the base has become *less positive* than the emitter. That's the same as saying it is *more negative* than the

emitter. The emitter-to-base bias has become 0.05 volt negative. (Call it emitter-base bias, not base-emitter bias. You want the emitter as the point of reference, so name it first.) An npn transistor with the base negative is reverse-biased. Collector current can't flow.

This should make clear that, even though you measure voltages with your voltmeter common lead connected to ground, the important thing is the voltage between elements of the transistor. In most transistor stages, your chief interest is the voltage between emitter and base; of secondary interest is the voltage between emitter and collector.

Suppose someone tells you the base voltage on one of these transistors has

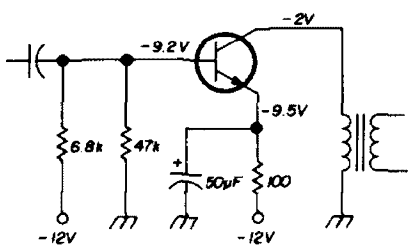



**fig. 2. Changing polarity of power supply doesn't alter circuit arrangement or operation.**

"gone up." What does that really mean? Usually he means the voltage is higher in the polarity shown on the schematic.

Consider the base voltage in **fig. 2**. It appears "lower" than the emitter voltage. Its value is less. Measured to ground, the base voltage is less negative than the emitter voltage. The important thing is this: being less negative, the base is *more positive* than the emitter. That provides forward bias for any npn transistor.

If the base voltage goes up—that is, if it goes further negative with respect to ground, as the voltmeter measures—the bias actually decreases. Say the meter measures −9.4 volts. The base has become more negative than it was. Looking from the standpoint of emitter-base bias, it tells you more if you say bias has become *less positive*. Forward bias is therefore reduced. Your voltmeter thus shows base voltage higher than before, but bias is less.

These are important relationships in transistor repair work. The simplest way to combat this seeming ambiguity is to quit using such vague notions as "up" and "down" for voltage measurements. Form the habit of thinking more negative or less negative, more positive or less positive.

## tests that reveal

At the repair bench you are usually concerned with a transistor in some piece of equipment. Tests you can make without unsoldering the transistor are the handiest.

There are three ways to evaluate a transistor in that circumstance. Two additional tests can be made if you unsoldered one or two transistor connections.

Finally, two quick test procedures evaluate a transistor outside the circuit. They are especially handy if you have a batch of unidentified transistors you want to check out. Even these tests can tell you more about transistor quality than you might expect.

## voltage measurements

Once you examine dc flow in transistor stages, you can figure out a lot from the voltages. If a voltage is wrong, deduction can tell you whether it's the transistor or something external.

Pretend the stage in **fig.3** is giving you

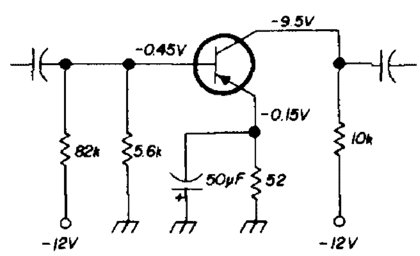



**fig 3. Pnp transistor in this basic amplifier works the same as npn; only change involves voltage polarities on the various elements.**

trouble. Your voltmeter tells you the base actually has −5 volts on it instead of the low −0.45 volt that's normal. Think out the possible causes.

Could be one of the base resistors is bad. But collector-base leakage in the

transistor is far more likely. You can verify by disconnecting the base lead of the transistor. If voltage on the open base lead is still highly negative, the transistor junction is leaky.

Or, in the same stage, suppose the emitter measures −0.9 volt. For some reason, more current than normal is flowing in the 52-ohm resistor the emitter voltage is measured across. The transistor is probably drawing too much current.

But is that due to overbias or a transistor defect? If base voltage has remained about the same, the trouble is likely in the transistor. You see, −0.9 volt at the emitter, with only −0.45 volt at the base, constitutes reverse emitter-base bias for this pnp transistor. That would reduce current through the transistor, not increase it—*unless* the transistor happens to be defective.

There are plenty of other examples of this kind of reasoning. Just remember which polarity of transistor you're dealing with and the likely effects of voltage changes. And don't forget to interpret voltage measurements in terms of their relation to each other and to the transistor itself.

The other two in-stage test ideas utilize a transistor's bias characteristic. For most transistors, zero and reverse bias cause zero collector current. A healthy forward bias assures significant collector current. These precepts of course apply only if the transistor is not defective.

The first test is for stages in which the transistor operates with forward bias. You can determine that from the schematic. Remember, forward bias is base-positive for an npn transistor and base-negative for a pnp.

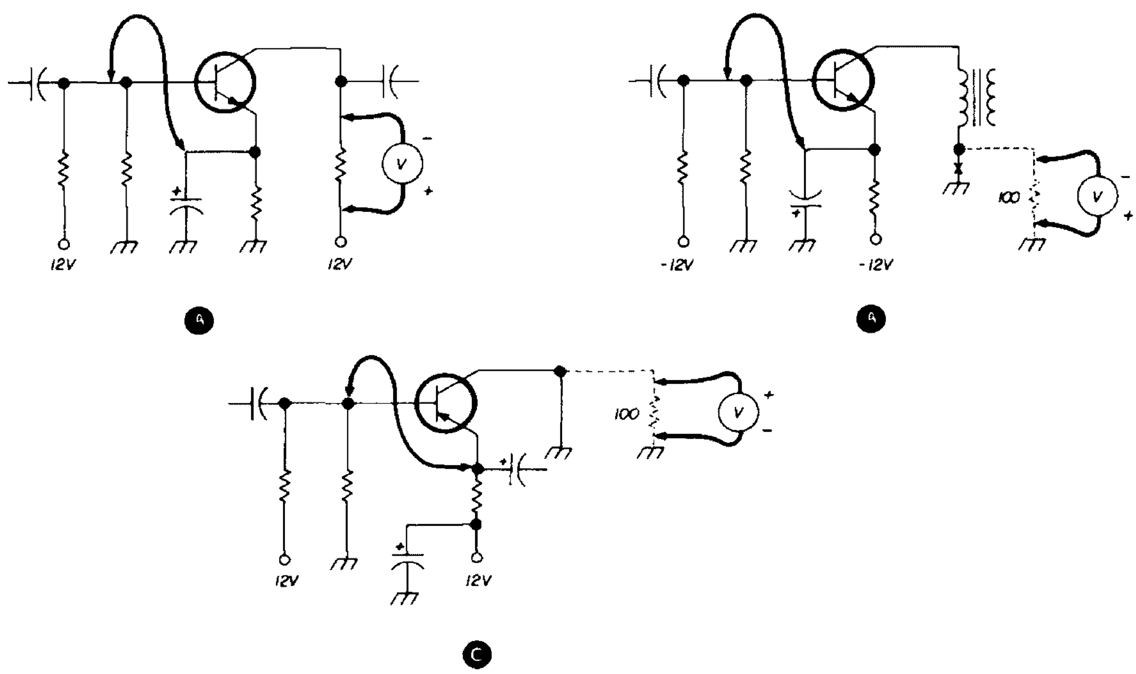



**fig. 4. Voltmeter connections in several amplifier stages for making bias-change operation tests. Idea is to eliminate bias on stages that normally use forward bias and add it to those that don't, while watching the change in collector current. You can add a resistor if the collector circuit doesn't have one.**

Connect your voltmeter at one of the points shown in **fig. 4**. Several possible connections are illustrated. If you need it, you can insert the 100-ohm resistor; its value won't bother the circuit much. Indirectly you are measuring collector current.

Notice the voltmeter reading. Then clip a shorting jumper between base and emitter. The voltmeter reading should drop to almost nothing. If it doesn't, the base isn't controlling collector current.

The second test is for stages where zero or reverse bias is normal. (The transistor may conduct, but probably during only a small portion of each signal

cycle, leaving an *average* or dc bias that is zero or reverse.) The voltmeter connections are the same as in **fig. 4**.

This time, instead of eliminating bias by shorting base to emitter, you apply a definite forward bias to base. Figure out from the schematic what would constitute forward bias for the transistor. Then somehow alter the bias to make it temporarily forward. The meter reading should take a definite move upward, signifying more collector current.

For instance, the npn transistor in **fig. 4A** has forward bias only when the base is more positive than the emitter. How do you make it more positive? One way is to reduce the value of the supply resistor, since it goes to a positive voltage source. Just bridge it with a low-enough resistance to make the base more positive than the emitter. If the transistor is working normally, the voltmeter shows more collector current.

In **fig. 4B** the basic supply scheme is different. But the transistor is still npn. Forward bias requires base to be more positive than emitter, same as always. But how can you make it that way? Just remember that more positive is the same as less negative. Bridge a lower resistance from base to ground, low enough to reduce the base voltage to a value less than at the emitter. Collector current goes up. If not, the transistor isn't responding as it should.

The transistor in **fig. 4C** is pnp. Forward bias demands a base more negative (less positive) than the emitter. It should by now be easy for you to figure how to make this base less positive. When you do, the voltmeter should register higher collector current.

## detecting abnormal leakage

Those tests let you know a transistor can control its collector current. That's the key factor. But there's another factor that can keep a transistor stage from performing up to par. You need a way to check *leakage*.

Basically, it's easy. Your voltmeter and soldering gun are the only equipment you need.

The leakage that can most upset stage operation is from collector to base. The collector junction of an operating transistor has a high reverse bias. If that junction lets "carriers" through in the wrong direction, transistor gain is poor.

To measure collector-base leakage, disconnect only the base lead of the transistor. Clip the voltmeter common lead to the emitter. Set the voltmeter as if you were measuring collector voltage. Touch the other test lead to the free end of the base lead. Voltage there should be almost nonexistent. Unwanted leakage lets current across the junction to the meter.

## testing out-of-circuit

If you have a transistor tester, fine. With a good one you can test transistors in or out of the stage faster than with the tests I've outlined here. But if you don't have one, you may often need these procedures.

Tests outside the stage are popular with hams. The basic instrument is your ohmmeter. There are two main purposes. One is identification. The other is evaluation.

Hams often pick up transistor "bargains." You get a handful of odd-lot transistors, often unmarked or marked in some way that means nothing to you. You may not know if they're pnp or npn. You might not even know which wires go to emitter, base, or collector. Here's how to settle these doubts.

An ohmmeter with 1.5 volts or less between the test leads is safest (measure with some other voltmeter). More voltage might pop a transistor junction. Also, notice which test lead has the positive voltage and which the negative; you'll need to know for these tests. Nowadays, it seems most ohmmeter batteries are connected with positive voltage on the common or black test lead.

Pick any two transistor wires. Clip the ohmmeter to them in first one direction and then the other. If you get no reading, try another pair, again measuring in both directions.

When you get a low ohms reading (150 or less), one of the ohmmeter leads is

clipped to the base wire. The way most transistors are arranged, it is the wire in the middle.

But you can make sure. Leave one ohmmeter lead clipped to the wire you think goes to the base. Move the other lead to the remaining transistor wire. If the ohmmeter reading is again low, the lead you didn't move is definitely clipped to the base. If not, the one you moved was.

You can now identify the transistor type. When you get low readings to both other elements with the positive ohmmeter lead connected to the base, you are testing an *npn* transistor. A pnp transistor gives low readings when the negative ohmmeter lead is clipped to the base.

You've identified the base, but you don't know which of the other two wires goes to the collector. There were clues in years past, but you can't trust the dots, stripes, and tabs on today's myriad of transistors. And basing diagrams aren't standard enough to help much either.

Start with the ohmmeter connected to show low resistance between the base and either of the other elements. You know which wire is base, so unclip that lead and move it to the other unidentified wire. The meter should read infinity, or open. If not, the transistor is defective.

Then click the range switch of your ohmmeter to higher scales until you see a slight downward meter deflection (something less than infinity). This usually happens on the Rx10k or Rx100k range. Next, reverse the two ohmmeter leads. The ohms reading will either go lower or return to the infinity end of the scale.

Connect the leads for the lower reading. Of course they are between emitter and collector. The negative ohmmeter lead is at the collector. This works for npn *or* pnp. Put a spot of paint or fingernail polish by the collector wire so you can identify it thereafter.

## leakage by ohmmeter

The tests you've already made tell you if a transistor is leaky or shorted. It's just a matter of interpreting.

When you've established the two low-resistance readings from the base, *notice* the readings in the reverse directions. If they're under 10k for either junction, there is too much leakage.

If you find low readings in both directions between any two leads, that junction is shorted. If a reading between two leads shows open both ways, even on the Rx100k scale, that junction is open.

A reading less than 10k from collector to emitter, in either direction, indicates too much leakage.

**Two-step method for identifying transistor type and base, collector and emitter connections. You need only your ohmmeter, but the transistor should be out of the circuit.**

### ohmmeter tests

**Step 1.** Find transistor lead that measures low R(150 ohms or less) to both other leads; that is the base lead.

| | | |
|---|---|---|
| If the ohmmeter *lead on the base* goes to the . . . | negative | positive |
| end of your ohmmeter battery, the transistor is . . . | PNP | NPN |

**Step 2.** Connect the ohmmeter for lowest R (above 10k) between the remaining transistor leads.

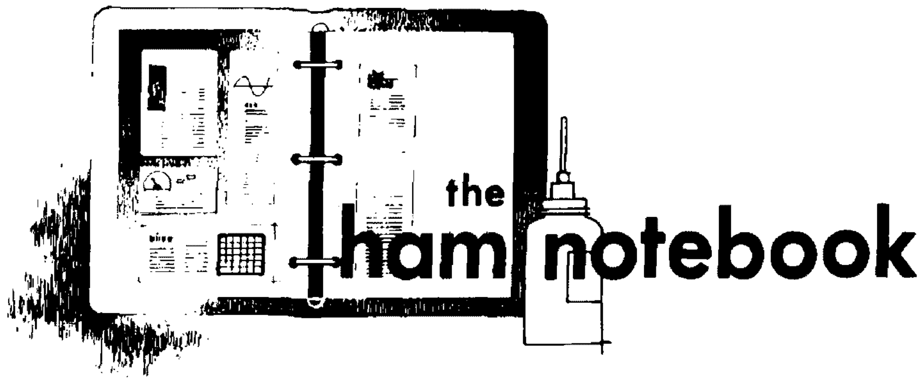
The negative ohmmeter lead identifies the collector

## next month

While your attention is on transistors, perhaps it's a good time to go into a complaint hams hear first from other hams. Ever been accused of putting out a distorted voice signal? It's easy for distortion to creep into speech amplifiers and you never know until someone hollers.

Usually the tests outlined this month spot any transistor defect that might cause distortion. But other parts in the stage can be responsible. And then there's always the oddball transistor that tests okay but goes ahead and creates distortion anyway. In my next column I'll tell some ways you can get rid of that kind of trouble.

**ham radio**




## surplus ic's

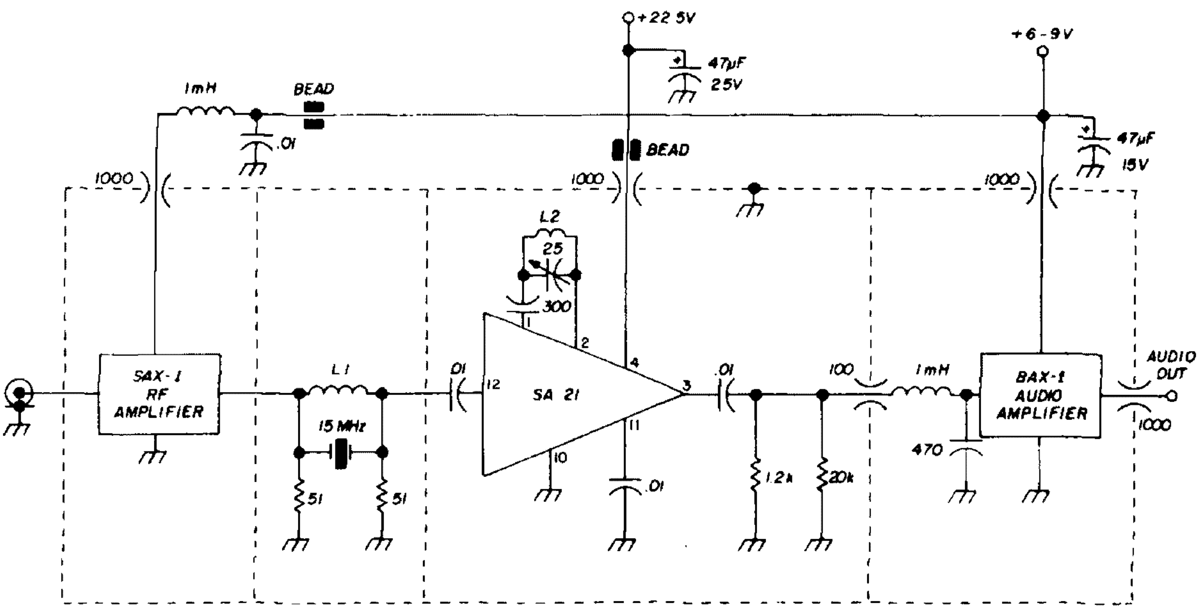
Now that surplus digital integrated circuits are so commonplace, more and more experimenters will be using them in their circuits. Unfortunately, most of these low-cost ICs use 14-lead flat packs—with leads spaced 0.05 inch apart. This small spacing and the natural fragility of the IC makes them extremely hard to solder into a circuit.

## simple wwv receiver

The simple fixed-tuned WWV receiver shown in **fig. 1** uses a crystal filter and an IC i-f amplifier designed by W6GXN (*ham radio*, November 1969, p. 24). Rf and audio amplification are provided by two inexpensive experimenter's modules available from the International Crystal Company. Selectivity with the crystal filter is excellent—on the order of 6 kHz—more than sufficient for reliable WWV reception.

Doug Pongrance, WA3JBN

*The SAX-1 transistor rf amplifier is $3.50; be sure to order the low kit (3 to 20 MHz) for this application. The BAX-1, a broadband untuned amplifier that is useful from 20 Hz to 150 MHz, is $3.75. Order from International Crystal Mfg. Co., Inc., 10 North Lee, Oklahoma City, Oklahoma 73102.




L1 10 to 18 μH (CTC X2060-4)

L2 3.3 μH. 30 turns no. 26 enameled on a T44-10 core

**fig. 1. Simple WWV receiver uses active i-f filter for selectivity. The SAX-1 and BAX-1 modules are available from the International Crystal Company.***

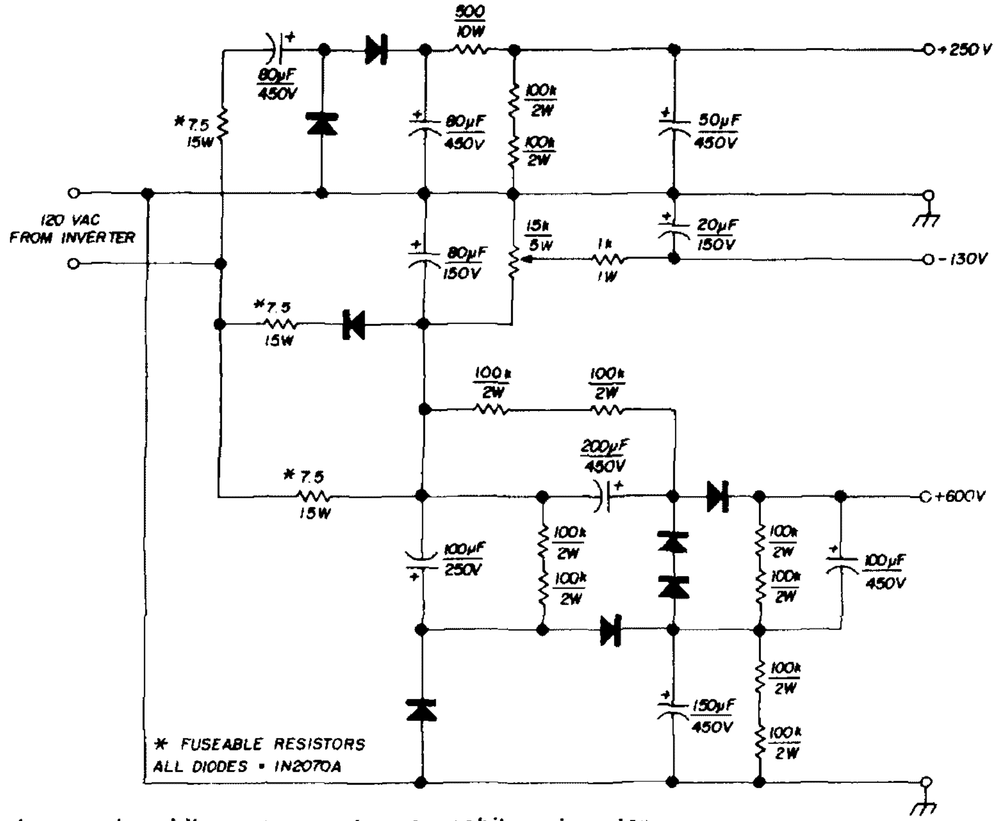



fig. 3. Low-cost mobile power supply uses mobile ac inverter.

However, I have found a simple method that works out very well. First of all, bend leads 2, 4, 6, 9, 11 and 13 down at the case at a 90° angle. Now all the leads are spaced 0.1 inch apart; much better for soldering and making circuit board connections. Bend the remaining leads down, out from the case an 1/8 inch or so.

Use the layout shown in fig. 2 for putting the ICs on a circuit board. I made a small template from a piece of aluminum, and attached it to the circuit board with a piece of Scotch tape and drilled

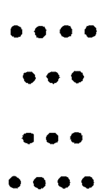


fig. 2. Full-size drill template for mounting 14-lead integrated circuits on printed-circuit board.

the holes in the board. With this much spacing it's not too difficult to draw circuit trails. You could accomplish approximately the same thing with perforated board with holes on 0.1 inch centers, but the problem here is connecting wires to the fragile IC leads. It's much easier and quicker with an etched circuit board.

Always check out the IC before you solder it into place. It's next to impossible to remove one of these 14-lead IC's from a printed-circuit board without destroying it. Flip-flops can be tested with the output from your 100 kHz calibrator. Run a lead from the output of the IC to the receiver antenna socket; if the flip-flop is working properly you should pick up calibrate signals at the 50 kHz points on the dial. Integrated-circuit gates can be given a cursory checkout with dc levels. For more involved testing you'll need a square-wave generator and an oscilloscope.

Nat Stinnette, W4AYV

## low-cost mobile supply

Recently I wanted to try mobile operation with my Heath HW-32A, but I didn't want to invest in an expensive mobile power supply. Since I had a dc-to-ac inverter on hand, I decided to build a transformerless voltage quadrupler and doubler circuit using the inverter as the ac source. The circuit is shown in fig. 3. Output voltages are +600 volts, +250 volts and -130 volt bias.

Henry Frink, W4GEG

## surplus relays

Surplus 24-volt dc relays are often available to the amateur experimenter at attractive prices. Most of these relays are well designed (and originally very expensive), but hams tend to steer clear of them since a 24-volt dc power supply is required. Not so. I've been using these relays, powered off the ac line, for many years; not even a transformer is needed, just a silicon diode, a resistor and a filter capacitor.

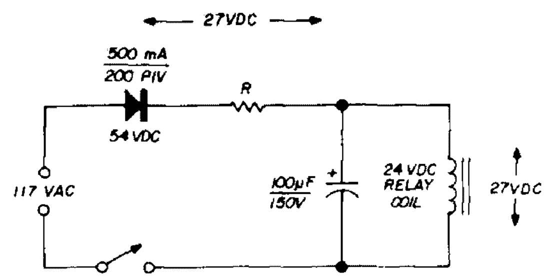
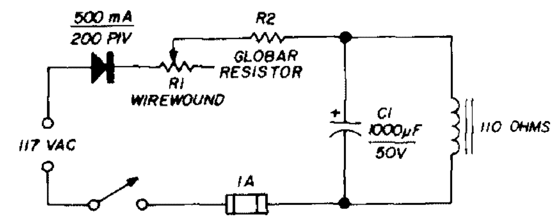



**fig. 4. Operating a 24 Vdc relay from the 117 Vac line. Resistor R is a 10-watt wirewound unit, the same resistance as the relay coil.**

Some of the late model hermetically-sealed relays have a relatively high resistance, but most of the open-frame types have a resistance in the range from 150 to 500 ohms. With the circuit shown in **fig. 4** the dc output from the diode is 54 volts. If the resistor is chosen with the same resistance as the relay coil, 27 Vdc will be impressed across the relay. Use a 10-watt wirewound resistor. The filter capacitor only has to be large enough to keep the relay from buzzing; generally 100 µF will be more than enough. An electrolytic rated at 150 volts costs only a few cents more than one with a 50-volt rating, and the higher rating is worthwhile in terms of trouble free operation.

You can obtain a small time delay (that is, the relay will hold in for a small time after the switch is turned off) of 1 to 2 seconds by simply increasing the filter capacitor to 500 to 1000 µF. For thisi application a 50 working-volt capacitor is suitable. For longer delay periods—up to 30 seconds—use a Globar resistor in series with the coil as shown in fig. 5. I tried two different types of Globar resistors, type FR-9 and FR-50. The FR-9 is used for replacement service in tv sets and is widely available. It works quite well with relays having 150 ohms coil resistance or less. The FR-50 works best with coils with greater than 150 ohms resistance. If you can't find an FR-50, it is interchangeable with type FR-100 and FS-800 (type numbers by Workman Associates, the distributor).

Time delays can be roughly predetermined from **fig.** 5. If a relay has too high resistance for a particular application, it can be lowered by the simple expedient of putting a composition resistor in parallel with the coil.




| time delay | R1 | |
|---|---|---|
| 4 - 5 sec | 500 | FR-9 |
| 8 - 10 sec | 500 | FR-50 |
| 15 - 20 sec | 250 | FR-50 |
| 20 - 30 sec | 250 | FR-9 |

**fig. 5. Time delays to expect with different Globar resistors (based on 110-ohm relay coil). Value of C1 can also be varied.**

The opposite effect—rapid pickup—can be obtained with the circuit shown in **fig.** 6. The two no. 313 pilot lamps have very low resistance when cold and this permits rapid relay pickup because of the initial over-voltage. Supply voltage is reduced to normal as soon as the lamp heats up and its resistance increases. An ordinary relay in this circuit, in parallel with 60 µF, was sufficiently responsive to follow keying at 20 words per minute.

**Neil Johnson, W2OLU**

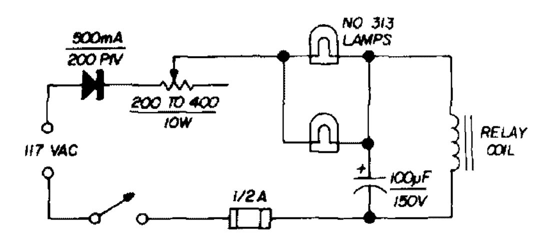



**fig. 6. Rapid armature pickup occurs in this circuit because of the low cold-resistance of the lamps, allowing an initial over-voltage.**

THE HAMMARLUND MFG. COMPANY
PERFORMANCE DATA FORM
LINEAR AMPLIFIER HXL-1

SERIAL NO. 33840268

DATE 3-11-70
BY JP / C.[illegible]

20 Ma. IDLE PLATE CURRENT (RELAY OPEN)
100 Ma. IDLE PLATE CURRENT (RELAY CLOSED)

NEUTRALIZATION

80M ✓ 40M ✓ 20M ✓ 15M ✓ 10M ✓

2500 VOLTS PLATE VOLTAGE ON STANDBY

USE USB 2 TONES OF EQUAL AMPLITUDE

PREPARED BY:

ROBERT MILLER, W4AEY
ENGINEERING LEADER

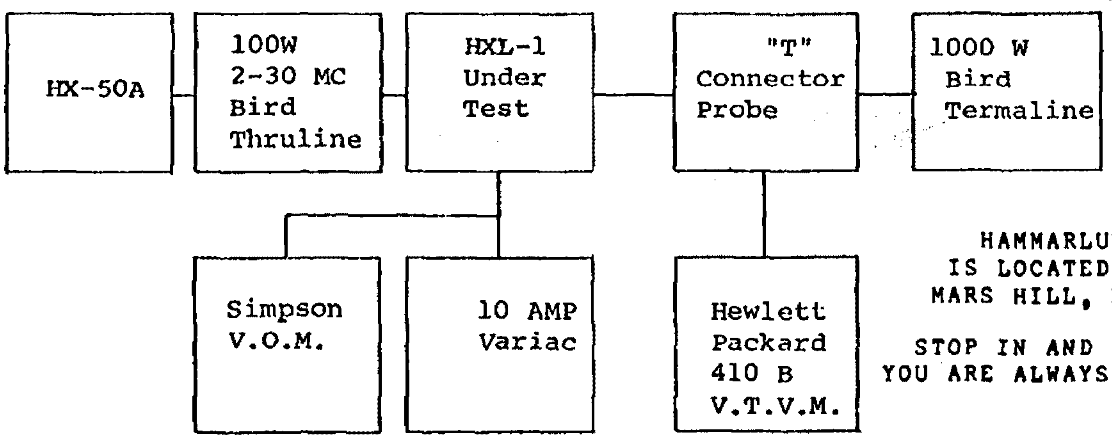
NOTE: ADJUST LINE VOLTAGE TO 110 VOLTS UNDER LOAD FOR POWER MEASUREMENTS.

| BAND FREQUENCY | RF DRIVE * * WATTS | PLATE VOLTAGE * * VOLTS | PLATE CURRENT * * MA | POWER OUTPUT * * WATTS | % EFFICIENCY * * | P.E.P. OUTPUT * * WATTS |
|---|---|---|---|---|---|---|
| 80M 3.8 MC | 100 | 1800 | 650 | 860 | 73.5 | |
| 40M 7.3MC | 100 | 1725 | 600 | 840 | 81.0 | |
| 20M 14.3MC | 100 | 1700 | 608 | 840 | 81.0 | |
| 15M 21.3MC | 100 | 1600 | 725 | 810 | 70.0 | |
| 10M 28.3MC | 95 | 1650 | 900 | 820 | 55.0 | |

EQUIPMENT SET-UP

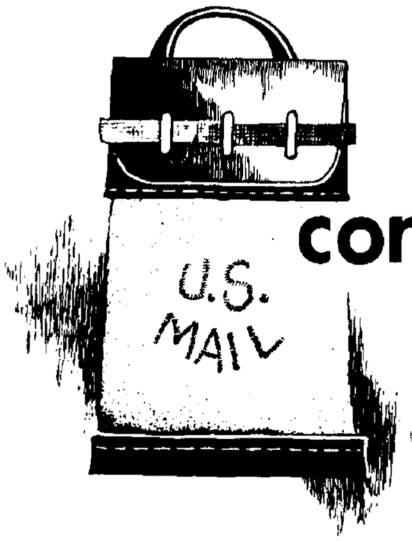
MULTIPLY x 2.5 FOR PEP

HELP YOURSELF!




HAMMARLUND
IS LOCATED IN
MARS HILL, N. C.

STOP IN AND VISIT.
YOU ARE ALWAYS WELCOME.




# comments

**Dear HR:**

Your comments in the January, 1970 editorial on QSL cards—a subject that will be with us to eternity! I am a keen-on-cards man, not necessarily juicy DX, but any cards, even from the next street. I QSL 100% via the bureau, and send a batch every other week, so no great delays on my part. My records show overall returns of 53%, with the U.S. returning *41%*. My operation is about 95% CW on the high-frequency bands, mainly 20 and 15, so I work plenty of Yanks. Only the Novices can be relied on to QSL, and they usually want air mail, which is too expensive when you work 5 or 6 in an evening.

My rig is a Heathkit HW-100 which gives an excellent account of itself with only 80 watts input (homebrew power supply provides only 480 volts to the PA). The antenna is an end-fed 200-foot wire about 23-feet high with three bends in it, so I sigh a little when I hear this "1 kW, 3-element beam up 50 feet" stuff coming over from W-land.

**George Rooney, G3MKH**
**Cheshire, England**

**Dear HR:**

I subscribed to *ham radio* without having seen a copy on the basis of the recommendation of a friend, and I was quite anxious to see what I had bought. I must say that I am delighted with the first two issues. The layout and readability are excellent, and the technical quality is easily on a par with the best engineering publications.

I am an electronics design engineer and brought the February issue to work. Several of the engineers in my group read the article on power supplies. One engineer, a digital type, said he really had not understood power supplies before.

Around here audio and radio are analog, and since I am the only one dealing with radio I am the "analog nut"—but they still wanted to read your "analog" magazine.

**W. G. Barrett, WA6ESV**
**Westminster, California**

## speech processing

**Dear HR:**

A small note concerning the speech clipper portion of Mr. Frank Jones' circuit for his phase-modulated two-meter transmitter in the February, 1970 issue: the clipping diodes, when connected to their source through a series resistor, do not act as clippers but perform a logarithmic conversion to the signal as per the article by Mr. Lee Richey, WA3FIY, on a speech processor in the January issue. The series resistor provides the constant current source needed for the log conversion. The lower value of resistance used by Jones in no way detracts from this function, other than loading the drive circuit more heavily, but allowing greater output. This is a characteristic of all common germanium diodes. (We usually use 4.7k series resistance driven by an emitter follower in this circuit.)

If large audio voltages are applied the waveform will appear to be clipped but not symmetrically, and the amount of drive needed cannot be obtained from the

circuit using the voltages shown. Thus, the audio would be logarithmically converted by this circuit and would appear to work very well with the fm circuitry shown but clipping would not normally occur.

It should also be noted in Mr. Richey's circuit that when the additional fet follower is used the gate and its bias resistor should be isolated from the diodes by a capacitor.

**W. Herbert Schieboid**
**Bloomfield Hills, Michigan**

## filters for speech clipping

**Dear HR:**

Although there appears to be a lot of literature available on rf speech clippers, all the articles I have been able to read completely neglect the question of filters, and this is where a major problem can easily arise.

Most technical literature simply states that the second filter (following the clipper) should be similar to the first (sideband-generation filter). This sounds all right until you find that no two filters are exactly the same.

For example, in my transmitter I use a Japanese-made Kokusai 455 kHz mechanical filter ($R_t$ = 10k). When I went to the agent to buy a second filter for my clipper, he showed me six, all with different frequency characteristics. I chose the one which was closest to the unit in my transmitter:

| | existing filter | new filter |
|---|---|---|
| center frequency | 455.01 kHz | 454.99 kHz |
| 6-dB bandwidth | 2.5 kHz | 2.44 kHz |
| 30-dB points | 456.62 kHz | 456.66 kHz |
| | 453.41 kHz | 453.51 kHz |
| 60-dB points· | +2.4 kHz | +2.35 kHz |
| | -2.1 kHz | -1.87 kHz |

The effect of adding the second filter was a loss of output and the elimination of a small band of voice frequencies.

No doubt there are expensive filters available which would be satisfactory, but I doubt if amateurs could afford them. However, the solution is simple. The second filter, while having the same center frequency as the first, should have a wider passband, and this condition is easily met on 455 kHz by using a 15k Kokusai mechanical filter. The one I have shows 6-dB bandwidth of 3.4 kHz and 60-dB bandwidth of 6.2 kHz. There is thus no detrimental effect on the signal itself, but the harmonics are eliminated and quality maintained.

Unless this question of filters is dealt with in technical discussions amateurs could be put to a lot of expense leading to unsatisfactory results when building rf speech clippers. This would be a great pity because rf clipping, properly used, gives terrific results, and all ssb exciters need it.

**Syd Hùdson, ZL1AFO**
**Auckland, New Zealand**

## magazine shredding

**Dear HR:**

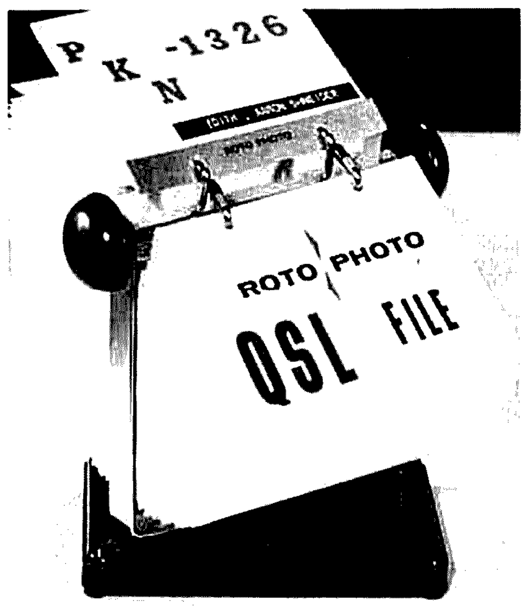
I hope that you will continue to adopt a sensible policy with respect to the names of articles. I appreciate informality as much as anyone, and indeed, our magazine holds informal presentation as a cornerstone of policy. But, I agree with E. L. Foster and believe the titles of articles ought to describe the contents, however briefly. If one uses E. L. Foster's magazine-shredding technique, one has often to re-label an article with its real title, rather than the entertaining but misleading one with which it was christened. I'd also like to suggest that it is not always necessary to use "transistorized" in a title, because of the very large number of transistorized items. It may even be useful one day to use "tube-operated" instead!

**Leo Gunther, VK7RG**
**Dynnyrne, Tasmania**

Leo is editor of *The Australian EEB*, an informal electronic experimenter's bulletin published about eight times a year. Subscriptions are $2.00 per year and well worth it. *The Australian EEB*, 32 Waterworks Road, Dynnryne, Tasmania 7005, Australia.

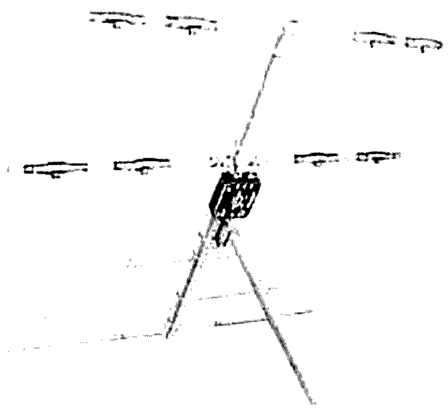
# new products

## rotary qsl file




This new rotary QSL file holds up to 500 QSL cards each in a clear plastic pocket. A turn of the knob brings each of the cards into view. Each QSL holder comes complete with 600 clear plastic pockets; refills are available to provide space for up to 500 cards. The rotary QSL holder takes up little room on the operating table, rotates on its base so it can be viewed from all sides, provides easy access and filing for your QSL cards and leaves your walls free for certificates and awards. Model CB-8-H rotary QSL file is $8 postpaid from M-B Products & Sales, 1917 Lowell Avenue, Chicago, Illinois 60639.

## beam antennas



Telrex has just announced a complete, new line of antennas under the brand name "Challenger." These antenna systems feature a balun-fed five-element Tri-band array for operation on 10, 15 and 20 meters using a single transmission line. The antennas have a peak power rating of 1 kW. Front-to-back ratio is rated at 28 dB. Other Challenger models range from ¾ through 80 meters. The entire Challenger line, as well as other Telrex antenna systems, are described in catalog PL70, available free by writing to Telrex Labs, Asbury Park, New Jersey 07712.

## multimeter

This new multimeter from Radio Shack is offered at a kit price but only needs a battery to be ready for use. It features 28 ranges that cover from 2 to 1200 Vdc full scale at 20,000 ohms per volt, 6 to 1200 Vac full scale at 10k ohms per volt, and 60 $\mu$A to 300 mA full scale. Four resistance ranges cover 1 ohm to 12 megohms, and a decibel scale covers -20 to +63 dB in five ranges.

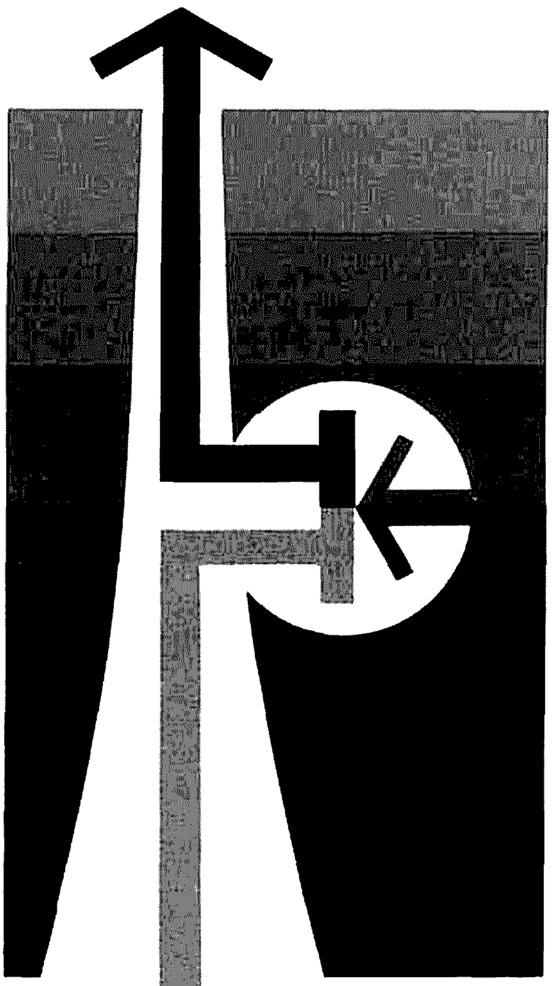
A jewelled meter movement and 1% precision resistors promise a long useful life. The mirrored scale prevents parallax errors when reading. The instrument is shipped complete with test leads, batteries and instructions. The model 22-022 multimeter is $14.95 at your local Radio Shack, or write to Radio Shack, 730 Commonwealth Avenue, Boston, Massachusetts 02215. An accessory leather case is available for $1.95.

75 cents

focus on communications technology . . .

# ham radio magazine

AUGUST 1970



## high-performance FILTER / PREAMPLIFIER for vhf/uhf receivers

this month

august 1970
volume 3, number 8

**staff**

James R. Fisk, W1DTY
editor

Nicholas D. Skeer, K1PSR
vhf editor
J. Jay O'Brien, W6GDO
fm editor

A. Norman Into, Jr., W1CCZ
Alfred Wilson, W6NIF
James A. Harvey, WA6IAK
associate editors

Jean Frey
art director
Wayne T. Pierce, K3SUK
cover

T. H. Tenney, Jr. W1NLB
publisher

John B. Morgan, K1RA
advertising manager

ham radio magazine

# contents



**offices**

Greenville, New Hampshire 03048
Telephone: 603-878-1441

**ham radio** magazine is published monthly by Communications Technology Inc. Greenville, New Hampshire 03048

Subscription rates, world wide: one year, $6.00, three years, $12.00. Second class postage paid at Greenville, N. H. 03048 and at additional mailing offices

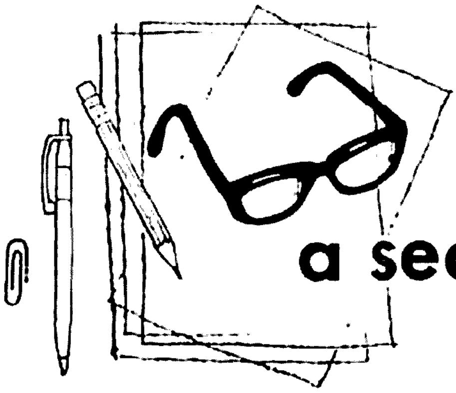
Foreign subscription agents: United Kingdom: Radio Society of Great Britain, 35 Doughty Street, London WC1, England.

All European countries: Eskil Persson, SM5CJP, Frotunagrand 1, 19400 Upplands Vasby, Sweden.


Printed by Wellesley Press, Inc. Wellesley, Massachusetts 02181, U.S.A.

**ham radio** is available to the blind and physically handicapped on magnetic tape from Science for the Blind, 221 Rock Hill Road, Bala Cynwyd, Pennsylvania 19440. Microfilm copies of current and back issues are available from University Microfilms, Ann Arbor, Michigan 48103.

Postmaster: Please send form 3579 to **ham radio** magazine, Greenville, New Hampshire 03048



# a second look

by jim fisk

LSI, or large-scale integration, is a term that is becoming more and more commonplace. An outgrowth of the transistor, and more lately, integrated circuits, LSI promises to benefit all of us by providing low-cost complex electronic systems for home and industry.

As production yields of less sophisticated devices such as linear integrated circuits have increased, so have the semiconductor manufacturer's efforts to put more and more electronic components on the same chip. Improved semiconductor technology, advanced manufacturing techniques, and new processes have resulted in LSI devices that defy imagination. The LSI chip shown in the photo, for example, contains 1,191 p-channel enhancement-mode mosfet transistors. This device, the Motorola MC1141, is a triple 66-bit shift register that is designed to operate over the frequency range from 10 kHz to 1 MHz, and is packaged in a long-lead version of the TO-100 metal can.

Much more complex circuits are in the works, and some devices are already on the market. In the new Boeing 747 jets, LSI has made it possible for passengers to have armrest control of their own lights, music and movie soundtracks. A small box containing an LSI chip is located on every third seat—this chip multiplexes the codes for the seat controls. Without LSI, 35 individual wires would have to be run to each seat.

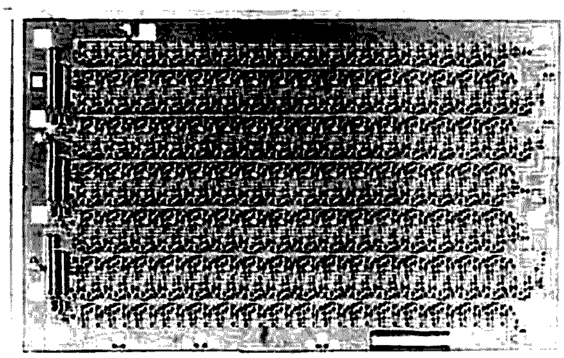
LSI is also being used in advanced computers and home calculators. Consider the computer built for NASA by RCA that is completely contained on a semiconductor chip one-seventh of an inch square, and performs all the arithmetical functions of a medium-sized computer. Or, how about the miniature calculator built by Canon (Japan), soon to be marketed in the United States, that adds, subtracts, multiplies and divides 12-digit numbers to four decimal places? The Canon calculator, which will be priced below $200 and weighs less than 2 pounds, contains three LSI chips, each an eighth of an inch square and containing 4,000 transistors.

Within a few years LSI may make possible such products as inexpensive home computers, trouble-free electronic controls for stoves and dishwashers, television sets no thicker than a picture frame, and telephones with built-in memories. With these products will come complex electronic instruments and equipment for the amateur that is now too complex or expensive to be practical.

If thinking miniature appeals to you, consider the prospect of 12 billion electronic circuits in a three-pount package—that's the *approximate complexity* and density of the human brain. With LSI, such an accomplishment may be a reality within the next decade or two.

**Jim Fisk, W1DTY**
**editor**

**Motorola MC1141 triple 66-bit shift register contains 1,191 mosfets on a single chip.**



# interdigital preamplifier and combine bandpass filter for vhf and uhf

High-performance filter/preamplification for vhf receivers that features low cross modulation, low noise and excellent unwanted-signal rejection

Bob Cooper, W5KHT, 6221 Norman Road, Oklahoma City, Oklahoma 73122

With the proliferation of vhf radio and television signals in virtually all metropolitan areas, weak-signal vhf reception that requires large antennas and extensive rf amplification has posed ever mounting problems to the receiver designer. Unfortunately, the requirements for weak-signal work are 180 degrees out of phase with an overload-proof receiving system.

You simply cannot erect a 16-dB antenna system for 50.1 MHz, follow it with a 20-dB preamplifier and install the whole works within five miles of a 100 kW channel-2 television transmitter—that is if you want the system to operate as if the 55.25 MHz signal was not on the air.

During the past five years or so, operation on 144 MHz has become equally challenging. In addition to the multiple transmitters located in the various public safety and business radio bands above 150 MHz, and the aircraft beacons and air-to-ground links in the 120 to 135 MHz range, we have high powered amateur repeaters within the band and MARS repeaters just outside. Also, activity on the 220-MHz band has always suffered in areas where channel 13 is in use, and channel 13 is allocated in practically all major metropolitan areas.

It all boils down to this: receiver desensitization—voltage overload of the receiver—is a serious problem that is slowly getting worse. As the occupancy of frequency allocations adjacent to our vhf amateur bands slowly builds up, the performance of our own communication systems becomes progressively more inadequate. Equally bothersome are the images and spurious beat products produced when two or more out-of-band carriers mix within our converters and generate new carrier products that fall within the amateur band we want to work.

These problems may be a constant annoyance (such as the beat product of a channel-3 video carrier at 61.25 MHz and a channel-12 video carrier at 205.25 MHz that produces buzz and hash on 144.00 MHz) or they may be intermittent. Carriers may add or subtract to produce the undesirable in-band product, or two mixing carriers may produce a spurious signal outside the band that mixes with a third carrier to produce an in-band birdie. The mathematical possibilities are endless and defy computation.

However, all these problem signals have one thing in common: they are created at some point in the receiver. The only way to cure the problem is to keep one or more of the mixing carriers out of the receiver stage where the mixing is taking place. It's common practice in many vhf converters to install a series-resonant circuit to trap undesired signals that cause spurious mixing products. However, this approach is not too effective, especially in strong-signal areas.

The complete solution is simple enough: make sure that the first stage of the vhf receiving system sees only that portion of the rf spectrum assigned to radio amateurs! However, solution is one thing, implementation another. Until W2CQH presented an amateur adaptation of interdigital and combline bandpass filters in *QST*[1,2] the usual amateur practice (and commercial as well) was to laboriously build a re-entrant cavity. The bandpass filter, when placed in front of the receiver, presents the receiver with an rf window a few MHz wide and severely attenuates all rf carriers that fall outside the window. The combline filters described by W2CQH offer simple construction, relatively narrow bandpass windows, steep bandpass skirts and extremely simple tuneup.

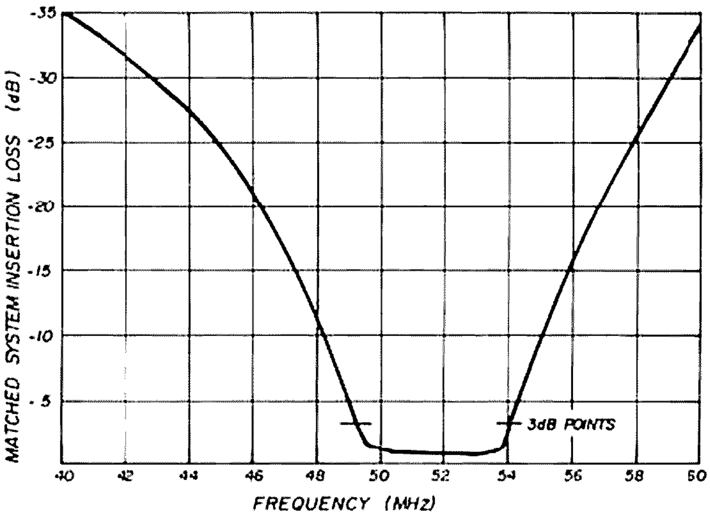
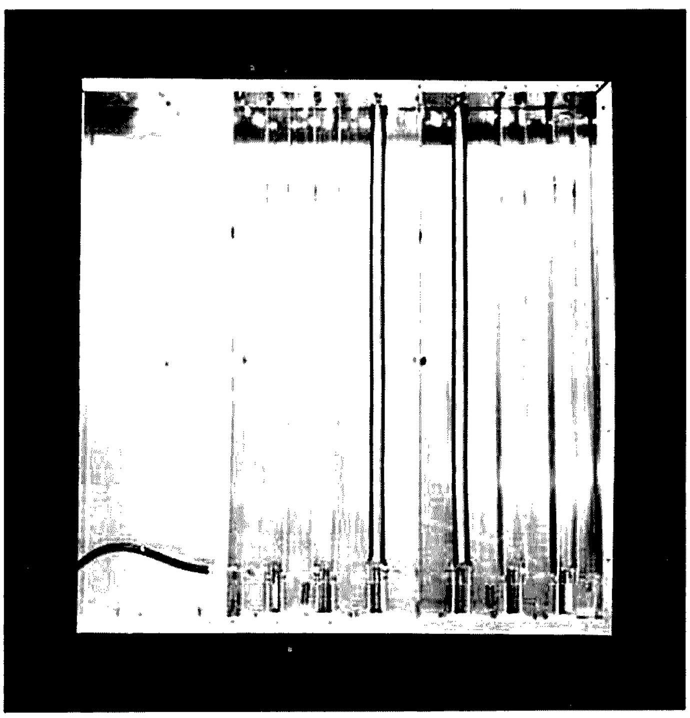



**fig. 1. Insertion loss of a 3-element combline filter centered on 52 MHz. The 3-dB bandwidth of this filter is 2.2 MHz.**

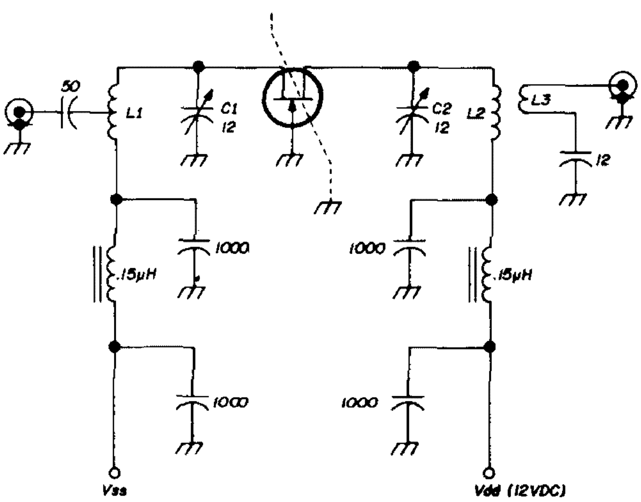
**Interdigital preamplifier for 50 to 100 MHz. Input is to the right, output on the left.**



Combline filters use strip lines for the L portion of the traditional LC network, loaded with variable capacitors to resonate the device to the desired center frequency. The 3-dB bandwidths of a three-element device are from 3 to 8 percent of the design center frequency. The bandpass window of a typical 3-element combline filter at 50 MHz is shown in fig. 1. Insertion loss of these filters is quite low; all the 3-element filters that I've built have had 1.0 dB or less insertion loss at the center of the passband.

## the preamplifier

The small-signal rf transistor has come of age as the radio spectrum has grown more and more crowded. The overload and cross-modulation problems associated with early transistors began to look less




L1 1.5" long no. 16 copper wire

L2 1.2" long no. 16 copper wire

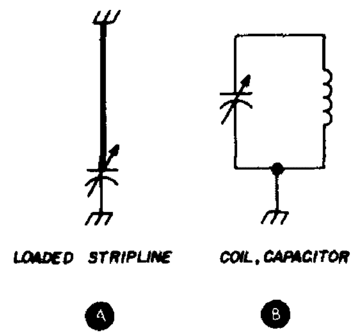
L3 2.0" long no. 16 copper, loosely coupled to L2 (¾" spacing)

**fig. 2. 450-MHz common-gate 2N5397 test circuit designed by Siliconix. In this configuration gain is 12 dB, noise figure, 4 dB.**

ferocious when the junction field-effect transistor was introduced. Early fets promised improved performance under taxing receiving situations that cause conventional transistors to quit, but early fet noise figures and gain made them less than useful above 100 MHz or so.

The 2N5397 jfet introduced by Siliconix* in 1968 was a major breakthrough for the receiver designer working above 250 MHz. While the 2N4416 created a boom in high-performance converters for the 144 and 220 bands (and occasionally 432 where it was derated), the 2N5397 offered super-simple circuitry. The 2N5397 grounded-gate circuit recommended by Siliconix for vhf applications, described in *ham radio* by K6HCP[3], results in as simple a preamplifier circuit as we could hope for.

**fig. 3. Capacitively-loaded stripline in A is the same as the parallel LC circuit in B.**




In describing the design parameters of the 2N5397, J. B. Compton notes[4] that ". . . this device has a typical operating transconductance more than double (the 2N4416) . . and this permits construction of common-source common-gate performance amplifiers that take advantage of the fet characteristics and give performance comparable to bipolar amplifiers over the frequency range from dc to 800 MHz."

The neutralization requirement of the common-source amplifier is eliminated with the common-gate arrangement. By eliminating neutralization you can design a tunable wide-range amplifier with amplification/bandpass characteristics limited only by the resonant characteristics of the input and output tuned circuits.

The common-gate approach is especially suitable because:

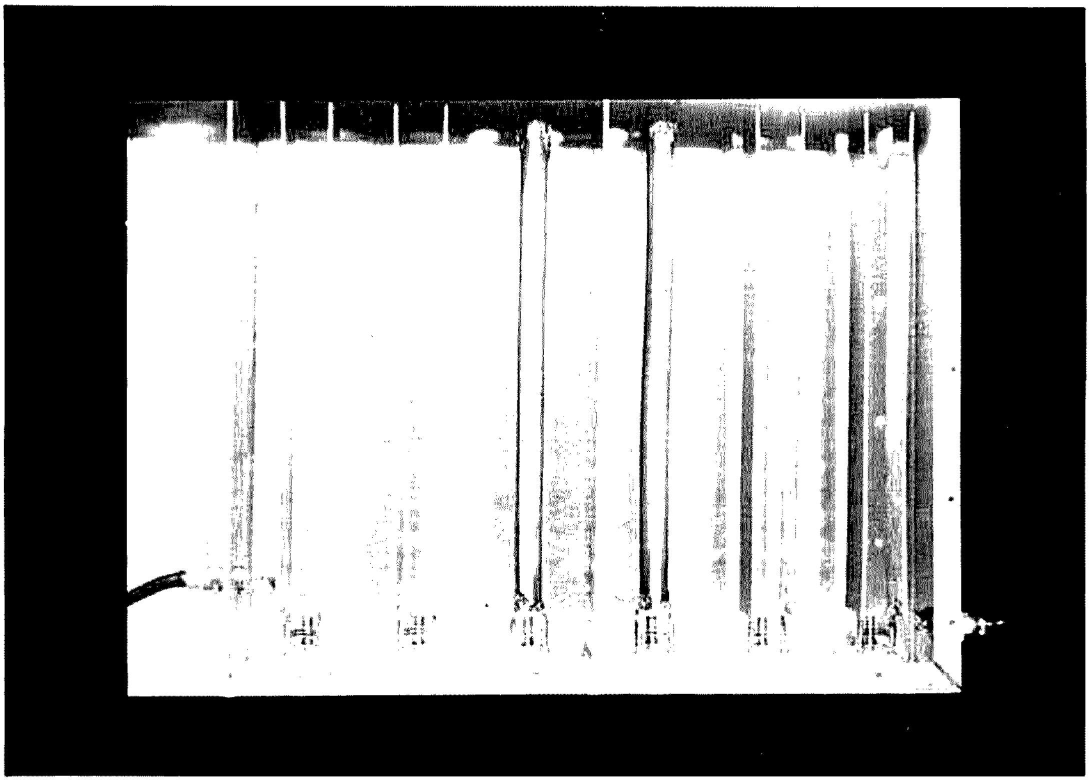
1. The input impedance is always lower than a common-source design,

*Siliconix, Inc., 1140 E. West Evelyn Avenue, Sunnyvale, California. The 2N5397 is available from W. Pat Fralia Company, Inc., P. O. Box 12625, Fort Worth, Texas 76116; $8.50 each.

although the output impedance is the same.

2. With no mismatch at the input a vhf noise figure of 2.0 dB or better is possible with the common-gate 2N5397. (Both K6HCP and I have found no difficulty in obtaining 1.5-dB noise figures at 144 MHz, and on 220 MHz I have never had a noise figure greater than 2.0 dB, nor greater than 1.0 dB at 50 MHz.)

without some loss in performance when compared to the common-source circuit. In 450-MHz test circuits developed by Siliconix for the 2N5397, a typical common-source amplifier produced 15-dB gain with a noise figure of 3.0 dB. A common-gate circuit developed a power gain of 12 dB with a noise figure of 4.0 dB. Slight retuning of the common-gate's input circuit lowers the noise figure to 3.3 dB and lowers the gain to 9 or 10 dB (fig. 2).



**Interdigital-filter/preamplifier for 120 to 240 MHz. The only differences between this unit and the 50-to-100 MHz design are the physical size and variable capacitors.**

3. The amount of amplifier-contributed cross-modulation and intermodulation distortion is proportional to the amplitude of the gate-source voltage. Since input power is proportional to input voltage and inversely proportional to input impedance, best amplifier performance is obtained with the lower impedance design of the common-gate circuit. A common-gate design offers approximately 10 dB more resistance to rf stage overload and cross modulation than a common-source arrangement.

However, the common-gate design is not

## filter/preamp combination

When a 3-section combline filter is cascaded with an identical filter, the 3-dB bandwidth points are essentially cut in half.* In other words, a 3-section

*The information presented here covers relatively narrow bandpass devices suitable for amateur applications. They are not suitable for television signal processing where 6-MHz bandwidths and 0.25-dB or less ripple is required. Both the units shown here and wider bandpass units suitable for television signal processing are the subject of patent applications filed by the author in March, 1970. Amateur construction of the units shown here for personal use will not violate the validity claims of the pending patent. editor.

5-MHz-wide combline filter at 50 MHz becomes a 2.5-MHz-wide filter at the output of the sixth section (two 3-section filters in cascade). A 5-MHz-wide filter at 144 MHz also becomes a 2.5-MHz bandpass in a two-unit system, while a typical 6-MHz bandpass on 220 is reduced to 3 MHz.

The factors that determine the width of the bandpass window are beyond the scope of this article. Suffice to say that the 3-dB bandpass can be changed within certain limits, both plus and minus, from approximately 10 MHz at 50 MHz to less than 500 kHz at 220 MHz. The information in this article should be suitable for most amateur applications.

## the filter

A single capacitively-loaded stripline represents a resonant circuit. Electrically, the loaded stripline shown in **fig. 3A** is identical to the LC circuit in **fig. 3B**. The efficiency of the loaded stripline is more dependent on the L/C ratio than the standard LC network where L consists of a coil. However, in the stripline system the Q can be made to track (with nearly constant Q factor) over a much wider range than the LC network.

Although W2CQH didn't mention it in his article[2] a typical 3-section combline filter is a one-octave device. That is, if the lowest resonant frequency of a given filter is 100 MHz with the loading capacitor fully meshed, the highest resonant frequency is approximately 200 MHz. By combining the flat characteristics of the combline filter with the flat performance characteristics of the 2N5397 common-gate amplifier you can build an amplifier/filter that exhibits essentially the same characteristics over a very wide band of frequencies.

**Fig. 4** shows how a pair of 3-section combline filters are combined with a single common-gate amplifier. In this circuit the third strip in the *first* 3-section filter becomes the tuned input circuit for the 2N5397, while the first strip in the *second* section becomes the output tuned circuit.

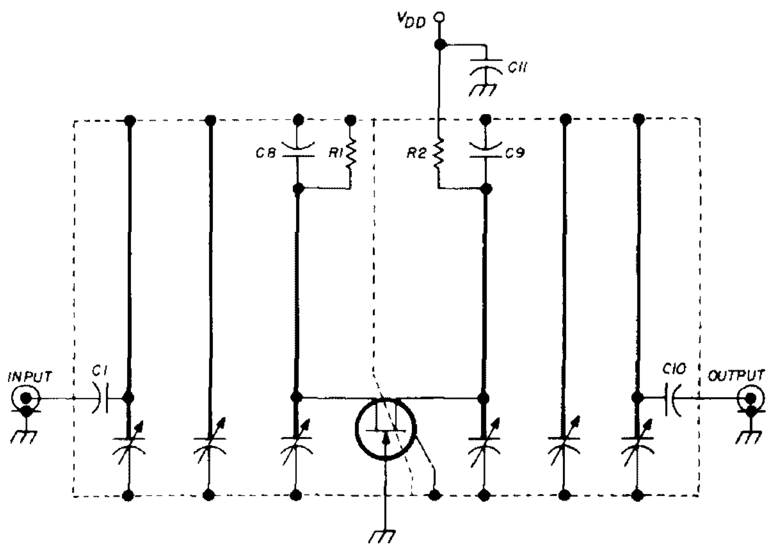



**fig. 4. equivalent circuit of the interdigital filter/preamplifier.**

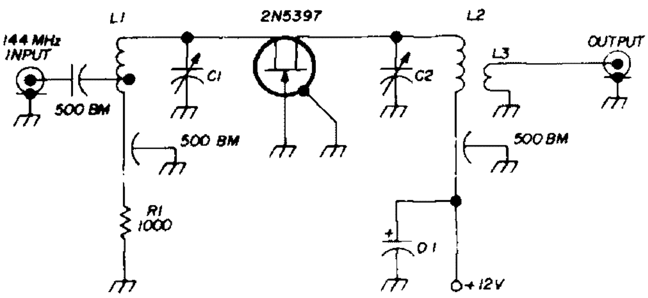
In the standard combline filter all strips are grounded at the cold end and tuned with the loading capacitors connected from the hot end to ground. In the filter/preamplifier, strips three and four are lifted above ground at the cold end. Strip three—the 2N5397 input circuit—is re-coupled back to ground through R1, a low value resistor chosen to provide the proper current drain for the fet. Strip number four is isolated from ground with a bypass capacitor (C9) and voltage fed through an isolating resistor. This circuit is nearly electrically

equivalent to K6HCP's 144-MHz preamplifier shown in **fig. 5**.

By combining the common-gate 2N5397 with the combline bandpass filter we have a low-noise moderate gain preamplifier with an input bandpass of 500 kHz or more.

## operational characteristics

The devices shown here were originally developed in an effort to solve an extremely difficult adjacent-channel tv interference problem at a catv antenna-




| | |
|---|---|
| **C1, C2** | **1.7-14.1 pF air variable** |
| **L1** | **5 turns no. 16, 3/8" diameter, 3/4" long, tapped at 1½ turns from cold end** |
| **L2** | **4¾ turns no. 16, 3/8" diameter, 1/2" long** |
| **L3** | **2 turns no. 20 insulated around cold end of L2** |

**fig. 5. Two-meter common-gate 2N5397 preamplifier described by K6HCP.**

receiving site. As a consultant in the catv field I had been called upon to make a 30- to 50-microvolt vhf tv signal appear free of noise and hash with broadcast quality color—not too tough until you consider the 100,000 microvolt adjacent-channel signal that came from a station less than five miles away!

Anyone who has fooled with fringe-area television is all too familiar with the broadness of tv frontends. The typical receiver has no more than 20-dB adjacent-channel rejection; the relationship between 50 microvolts and 100,000 microvolts is on the order of 66 dB. The problem was solved with an interdigital preamplifier similar to the ones shown here.

The common-gate 2N5397 preamplifier used in the filters operates just as

**table 1. Operational characteristics of 6-strip interdigital-preamplifiers suitable for 50, 144 and 220 MHz.**

| frequency (MHz) | 3-dB bandwidth (MHz) | 20-dB bandwidth (MHz) | gain (dB) | noise figure (dB) |
|---|---|---|---|---|
| 50 | 0.5 | ±4 | 10 | <1.0 |
| 144 | 1.0 | ±5 | 12 | 1.5 |
| 220 | 1.5 | ±6 | 12 | 2.0 |

expected. Stability is excellent. The unit tunes evenly with no instability from the lowest resonant frequency to the highest. Intermodulation and cross modulation within the preamp is as good or better than any solid-state device you can use. The noise figure of this device, as measured on working models, is just under 1.0 dB at 50 MHz, increasing to 2 dB on 220 MHz and rising to 4 dB at 450 MHz and 6 dB at 800; a noise-figure curve is plotted in **fig. 6**.

The gain of a single stage of amplification depends a great deal upon the L/C ratio of the striplines and load capacitors, as well as coupling between striplines and the match to the transistor. In the early stages of development I built nearly a dozen filter/preamplifiers to determine the proper balance of L and C. The performance characteristics were carefully measured as L and C were juggled,

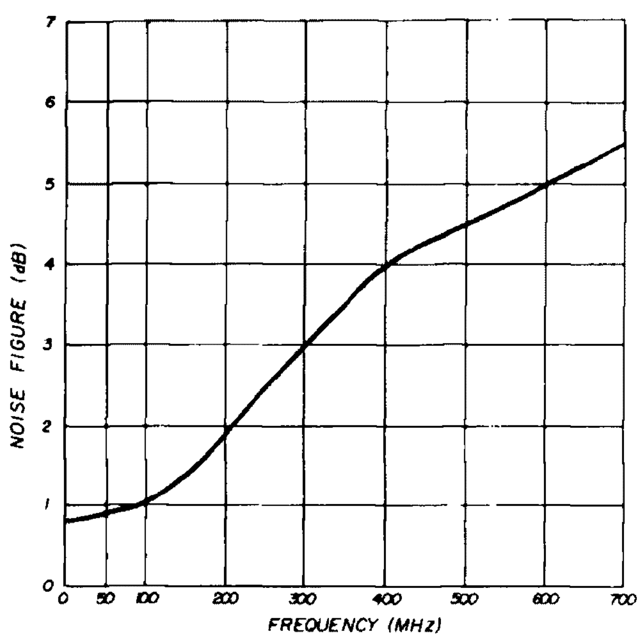
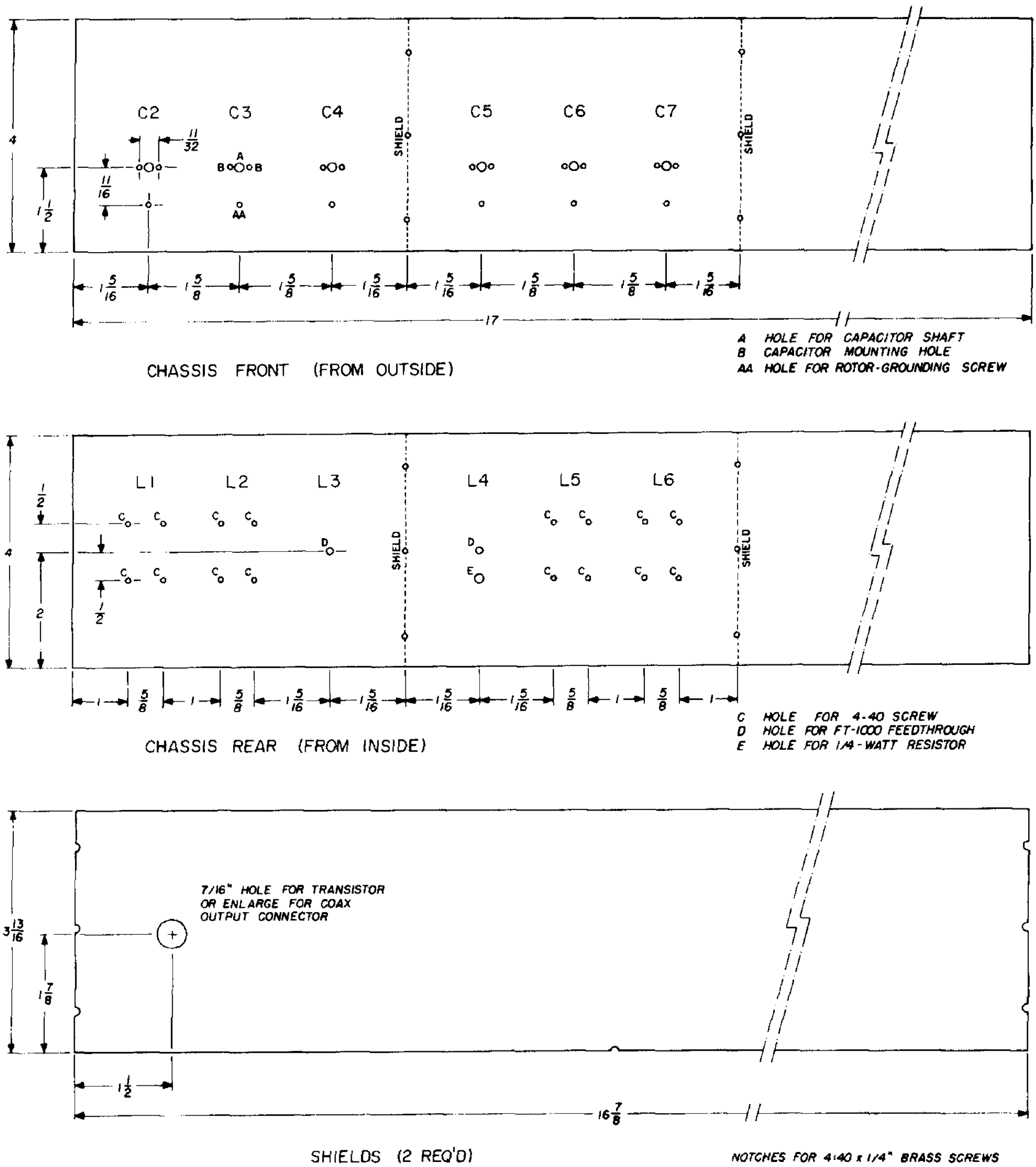



**fig. 6. Noise-figure curve for the grounded-gate 2N5397 preamplifier.**

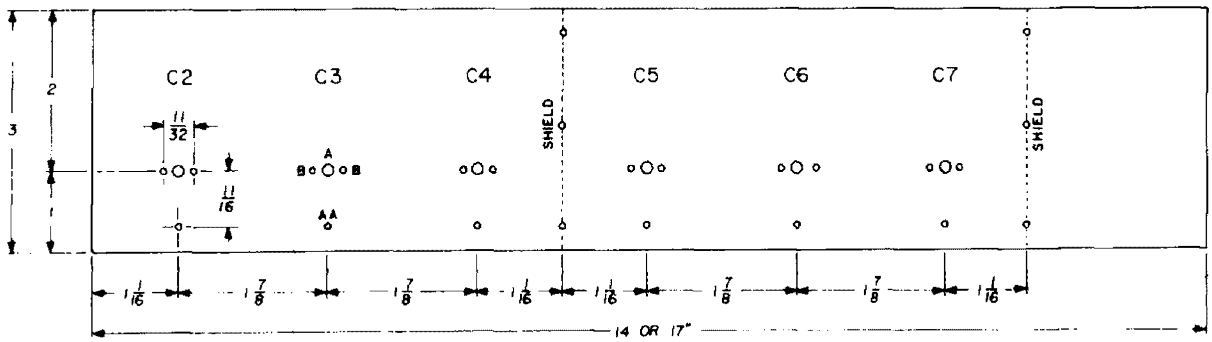



| | | | |
|---|---|---|---|
| **chassis** | **17 x 17 x 4" (BUD AC-1431)** | **shields** | **16-7/8 x 3-13/16"; cut from double-sided copper-clad printed-circuit board (Kepro P2-1212G)** |
| **cover** | **17" x 17" (BUD BPA-1597)** | | |

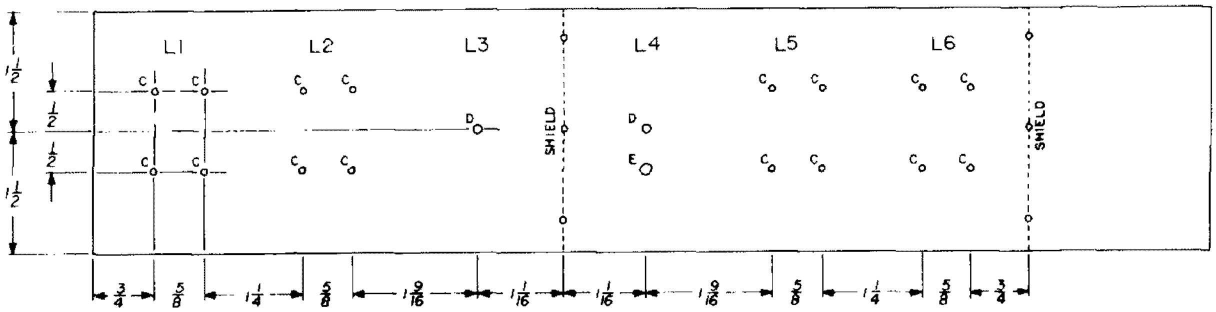
**fig. 7. Construction details of an interdigital filter/preamplifier that tunes from 50 to 100 MHz. Striplines are spliced as described in the text. Hole for input coax connector is centered, 1½" from bottom of chassis.**

strip mounting and strip-to-strip spacing varied, and input/output coupling changed. As in most repeatable designs, predictable patterns developed that provided the basis for design formulas.

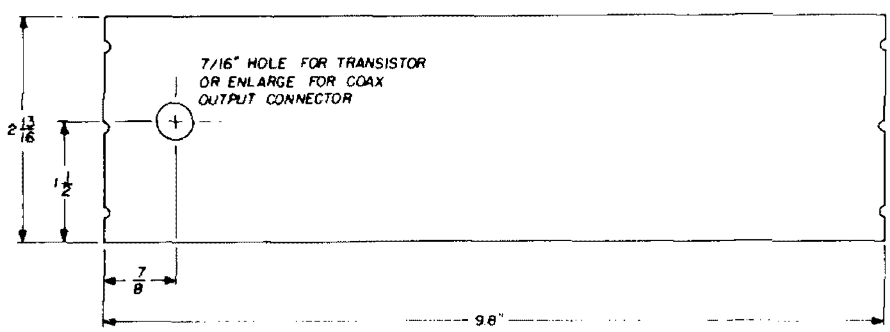
The information presented in this article will be limited to units that are especially suitable for the vhf amateur bands. The performance characteristics of 6-strip interdigital-preamplifiers for 50, 144 and 220 MHz are shown in **table 1**. Two identical interdigital-preamplifiers may be cascaded for any of the bands. This results in narrower 3-dB bandwidths,




CHASSIS FRONT (FROM OUTSIDE)




CHASSIS REAR (FROM INSIDE)




SHIELDS (2 REQ'D)

| | |
|---|---|
| **chassis** | **10 x 14 x 3" (BUD AC-414) or 10 x 17" (BUD AC-416). Required interior dimensions are 10 x 11¾", a non-standard chassis size; required size is provided by an interior end plate** |
| **cover** | **10 x 14" (BUD BPA-1524) or 10" x 17" (BUD BPA-** |
| **shields** | **9.8 x 2-13/16"; cut from double sided copper-clad printed-circuit board (Kepro P2-1212G)** |

**fig. 8. Construction details of an interdigital filter/preamplifier that tunes from 100 to 200 MHz. To raise the resonant frequency to 120 to 240 MHz remove one plate from C2 and C7; remove 2 plates from C3 and C6; remove 2 stator plates and 3 rotor plates from C4; and remove 3 stator plates and 3 rotor plates from C5. Hole for input coax connector is centered, 1-1/16" from bottom of chassis.**

considerably steeper 20-dB bandwidth skirts, and twice the gain of a single unit with the same noise figure.

## construction

Two versions of the interdigital-preamplifier are shown; one tunes from 50 to 100 MHz and the other covers from 120 to 240 MHz. Both are based on standard BUD aluminum chassis with bottom plates. The striplines are probably a little different than anything you've seen before—they consist of two pieces of copper-clad printed-circuit board, sup-

ported on the variable capacitor at the hot end and held by two ¼-inch 4-40 brass machine screws at the other.

Unlike the W2CQH combline, the striplines in the interdigital-preamplifier hang vertically. This effectively gives you twice as much surface area for a given resonator length and directly affects the efficiency of the LC networks, particularly at the lower operating frequencies. Striplines 1, 2, 5 and 6 use this technique; striplines 3 and 4, because of impedance matching to the 2N5397, use equivalent lengths of copper tubing. This design approach is several dB more efficient than all copper-tubing resonators, single horizontally suspended striplines or combinations of horizontally suspended strip and copper tubing.

The 2N5397 input circuit is shielded from the output circuit by a piece of double-sided copper-clad circuit board that runs the full width and full heighth of the chassis (see **fig. 7**). The 2N5397 is mounted in a 7/16-inch hole drilled in the shield; the case lead of the fet is soldered to the side of the shield facing stripline number 3, the gate lead is soldered to the side of the shield facing stripline number 4, the source lead connects directly to a pigtail from C3, and the drain lead is soldered to a pigtail on C4.

Input coupling to stripline number 1 and output coupling from stripline number 6 is accomplished with ceramic capacitors; values are given in **fig. 10** for either 52- or 75-ohm operation.

Striplines 3 and 4 are supported by Centralab FT-series feedthrough capacitors. Sprague BH-series stud-bypass capacitors could be used here, but construction is more difficult because of their small size. Stripline number 3 is resistance

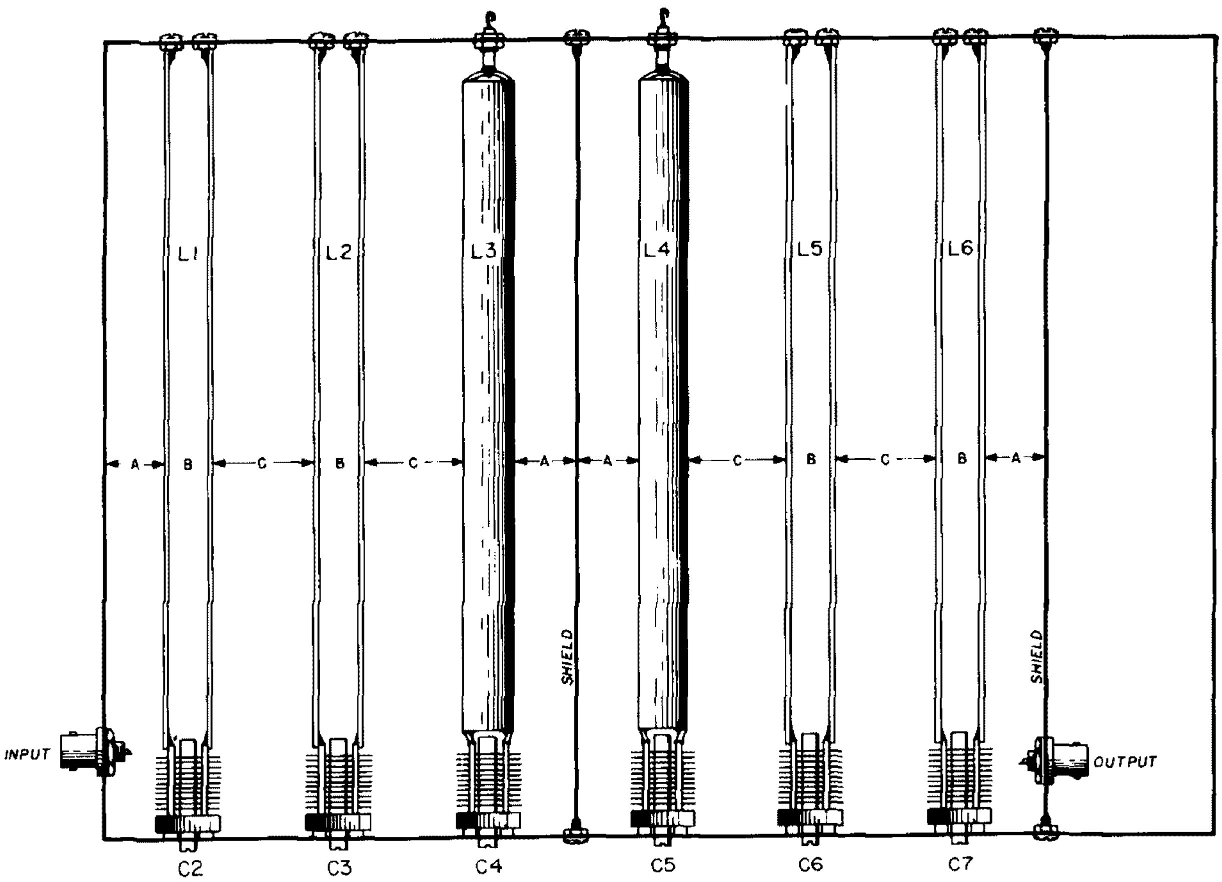



**fig. 9. Mechanical layout of the filter/preamplifier. Dimensions are determined by frequency range as shown below:**

| frequency (MHz) | A | B | C |
|---|---|---|---|
| 50-100 | 1-1/32" | 5/8" | 1" |
| 100-200 | 3/4" | 5/8" | 1-1/4" |

coupled to ground through a resistor that is chosen to allow the 2N5397 to draw 5 mA; typical values range from 100 to 470 ohms with 330 to 390 ohms quite common. Stripline number 4 is resistance coupled to the power supply through a 100-ohm resistor with additional bypassing on the outside of the chassis through a 0.1 $\mu$F ceramic capacitor.

Part of the BUD chassis is not used in both models of the filter/preamplifier shown here. In addition to the shield between striplines 3 and 4, another partition is installed after stripline number 6. This serves as the outer wall and holds the output coaxial connector.

## how to build it

Before starting construction the chassis should be laid out with each mounting hole center punched so each hole is drilled where it should be. Small errors in strip-to-strip spacing are tolerable, but anything over 1/16 inch is *not* small.

Sheet-metal screws are used to fasten the top plate to the chassis. It's manda-

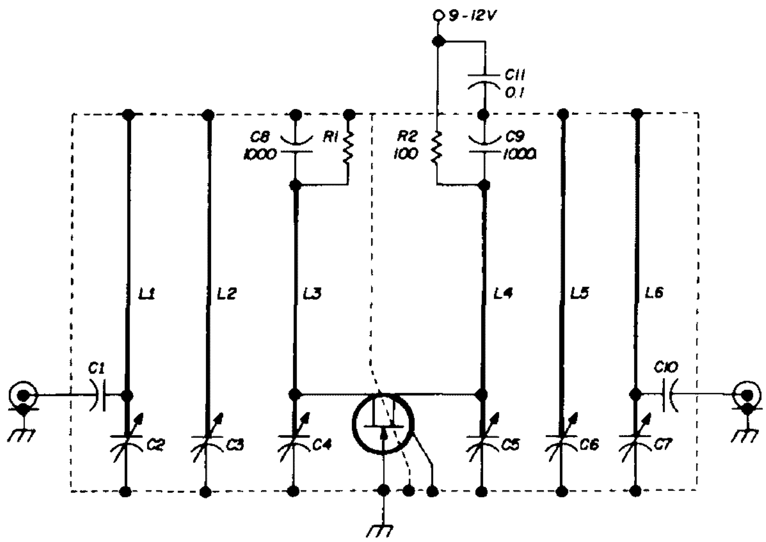



fig. 10. Schematic diagram for the interdigital filter/preamplifier.

| | |
|---|---|
| C1,C10 | 30-pF ceramic (75 ohms): 35—40-pF (50 ohms) |
| C2,C3,C4, C5,C6,C7 | 5—100-pF air variable (Calectro A1-226) |
| C8,C9 | Centralab FT-1000 feed-through capacitors |
| C11 | 0.1-$\mu$F ceramic, 50 volts working |
| L1,L2, L5,L6 | 8 strips (2 per stripline) cut from 1 oz. (0.00135) single-side copper-clad printed-circuit board (Kepro P1-1212G). Each strip 15¼" long, 1½" wide |
| L3,L4 | 5/8" OD copper tubing, 14¾" long |
| R1 | 100 to 470 ohms (see text) |
| R2 | 100 ohms, ¼ watt |

Components for 50 to 100 MHz unit

| | |
|---|---|
| C1,C10 | 6.8-pF cermaic (75 ohms); 10-pF ceramic (50 ohms) |
| C2,C3,C4, C5,C6,C7 | 2.9—30-pf air variable (Calectro AL-225) |
| C8,C9 | Centralab FT-1000 feed-through capacitors |
| C11 | 0.1-$\mu$F ceramic, 50 volts working |
| L1,L2, L5,L6 | 8 strips (2 per stripline) cut from 1 oz. (0.00135) single-side copper-clad printed-circuit board (Kepro P1-1212G). Each strip 8.9"long, 1.5" wide. |
| L3,L4 | 5/8" OD copper tubing, 8.4" long |
| R1 | 100 to 470 ohms (see text) |
| R2 | 100 ohms, ¼ watt |

Components for 100 to 200 MHz unit

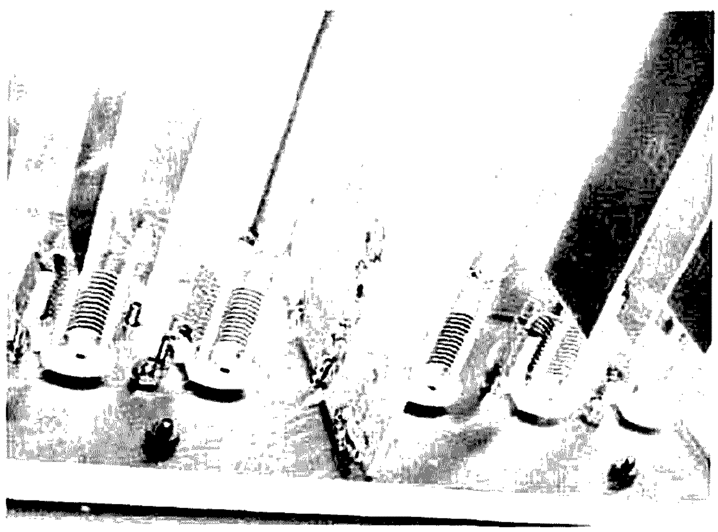
tory that this plate fits very tightly because, as well as a shield, it serves as a ground plane for the striplines. The sheet-metal screws are located at the end of each strip (on the cold end) and directly above the tuning capacitor (on the hot end). If you skimp on the number of screws you'll have problems with erratic tuning and low gain.

The striplines are centered between the top and bottom plates of the chassis. On the capacitor end of the resonator, the stripline is soldered to the capacitor stud with the copper foil facing in (all striplines are single-sided copper and mount on the outside of the studs so copper faces copper). At the cold end of the stripline brass grounding screws are located inside the strips directly opposite the capacitor studs. The copper-clad strips must run exactly parallel to each other or efficiency will drop.

The input and output ceramic capacitors must have short leads. Small holes drilled just above the center line on the input and output striplines allow you to feed the capacitor pigtail *through* the strip and solder it to the copper side.

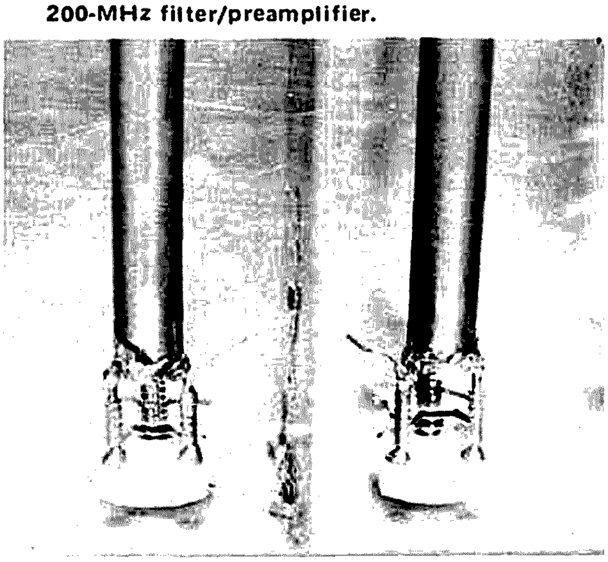
The two shields are made from double-sided copper-clad printed-circuit board and are cut to form a snug fit top and bottom, with small clearance on either end. Brass 4-40 machine screws, ¼-inch long, are mounted in the chassis wall at either end of the shield, and the shield is soldered to them. Make sure the brass screws are soldered to *both sides* of the shield. Small notches in the board allow the screws to seat snugly into the shield.

In the 50- to 100-MHz model of the filter/preamplifier an additional brass screw is installed through the chassis in the exact center of the shield. This holds the center of the long shield in place and helps to maintain the proper location of the shield in the long chassis.



**Mounting of center shield and striplines (50 to 100 MHz unit).**

**Transistor installation in 100 to 200-MHz filter/preamplifier.**

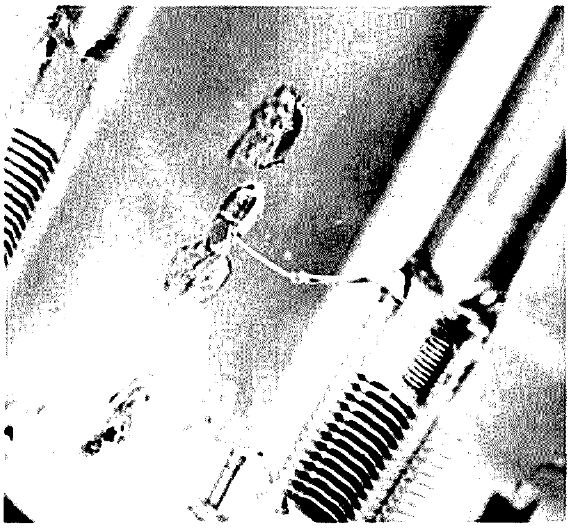


Kepro printed-circuit board is available in pieces up to 12 inches square; one 12 x 12 inch section will furnish 8 strips, 12 x 1½ inches. For the 50- to 100-MHz filter, the required 15¼ inch length is provided by splicing a 3¼-inch piece from another printed-circuit board. To bridge the splice joint, use thin strips of copper, 1¼ inches long, sweating the printed-circuit board and copper strip together with a 40-watt soldering iron. Be sure the 3¼-inch section joins the 12-inch strip flush on so the resulting 15¼-inch strip has no bends or kinks.

Striplines 3 and 4 are made from common copper tubing. Make sure the tubing is straight—you'll have to straighten the pieces you need carefully since it's usually sold in rolls. The hot end of the tubing fits nicely onto the capacitor studs. The cold end of the tubing is

attached to the mounting stud (FT feedthrough capacitor) with a short loop of number 16 or 18 copper wire that is soldered across the end of the tubing; the wire is soldered to the FT capacitor.

The rotors of the Calectro variable capacitor are not automatically grounded when the capacitors are mounted, so a ½-inch 4-40 brass machine screw is used to solder the rotor tab to ground. (Calectro variable capacitors are commonly



**Closeup of 2N5397 position in the center shield.**

available at the Calectro displays found at most electronic parts distributors.)

When you get down to actually building the filter/preamplifier it is recommended that you follow the step-by-step instructions given below.

1. Mark and drill all holes. First lay out the top plate, center-punch the holes and drill holes for no. 6 sheet-metal screws. Then put the top plate in position over the chassis and drill corresponding holes in the chassis lips using the top plate as a drill template.

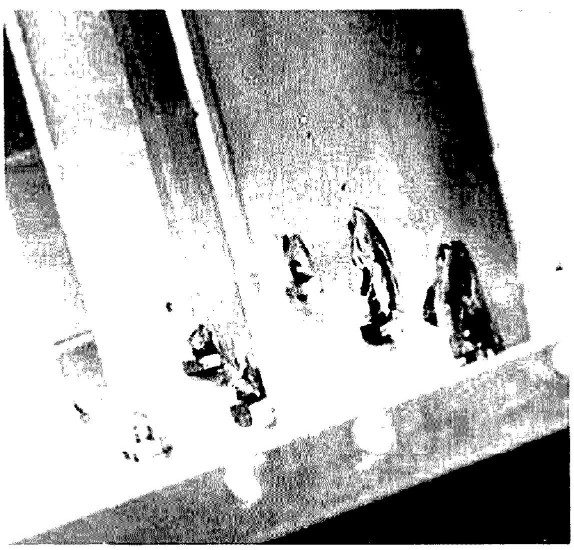
2. Install all of the 4-40 brass screws.

3. Install both shield sections. (If you wait until after you install the other components you won't be able to get them into the chassis.)

4. Install the variable tuning capacitors, grounding the rotor tab to the ½-inch 4-40 brass screws provided for this purpose. Install C8 and C9, the two FT-1000 feedthrough capacitors.

5. Install both coax connectors and the input and output coupling capacitors (C1 and C10). The stripline end is not connected yet.

6. Install the striplines—two per capacitor—starting with lines nearest the end of the chassis. Use a 25- to 40-watt soldering iron when installing the striplines because very little heat is required with the copper-clad strips. Solder the capacitor end first, holding the other end in position so that the weight of the stripline is not on the newly soldered capacitor stud. (If you mark the center lines of the strip you can align the centerline with the stud.) The cold ends of the striplines are held snugly by friction against the 4-40 grounding-screw nuts. Remember that for top performance the copper-clad strips must be parallel *and* centered between the chassis top and bottom ground planes.



**Cold ends of striplines are soldered to 4-40 screws.**

7. Now install striplines 3 and 4. You'll need a 150-watt iron (or more) to solder the copper tubing. Don't rest the weight of the tubing on the capacitor studs—block the tubing up until both ends are firmly in place.

8. Install R1—try 330 ohms to start—and install R2. Resistor R2 feeds through a small hole in the cold end of the chassis just under the FT bypass capacitor.

9. Solder the loose ends of the input and output coupling capacitors to striplines 1 and 6, pushing the pigtail through the small hole in the strip and soldering it to the copper foil.

10. Carefully install the field-effect transistor with the tab pointing toward the input side of the filter. Solder the case lead to the input side of the shield and the gate to the output side of the shield. The transistor leads can be preformed as shown in **fig. 11**; the source lead will point toward C4 (stripline 3), and the drain lead will point toward C5 (stripline 4). Short pigtails of no. 18 copper wire are soldered to the variable capacitor studs, and then to the free drain and source leads. When soldering the transistor leads use heat sinks and no more than a 40-watt soldering iron.

11. Install the power supply decoupling capacitor C11 *outside* the chassis.

## construction tips

If you are interested in duplicating the units shown in this article, there are a few tips that are worth passing along. If you experience problems with gain, and are satisfied that you have used proper layout and construction, remember that output is affected by power supply decoupling, output circuit loading (the device won't work well if too heavily loaded), and saturation. Power gain drops rapidly when the transistor is saturated—the 1-dB gain-compression point at 450 MHz, for example, is coincident with 1-mW input.

In addition, any inductance in the gate lead will severly affect performance. This indicates that the shield between stripline 3 and 4 *must* be well grounded to the chassis. If the side of the shield where the gate lead is soldered is floating above ground it will introduce considerable inductance to the gate circuit.

If any performance problems develop check your construction. The spacing between resonators, between the top and bottom plates, and the shield between striplines 3 and 4 *must* be uniform. The layout shown here was developed for optimum performance; if you deviate much, or use sloppy construction, gain will quickly drop off to 2 to 3 dB. Use extreme care when laying out the component mounting holes; variances of 1/16th inch will deteriorate performance.

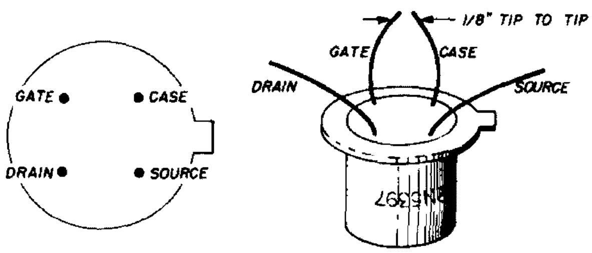



**fig. 11. Method of pre-forming the leads to the 2N5397 fet before installing it into the preamplifier.**

## tuneup

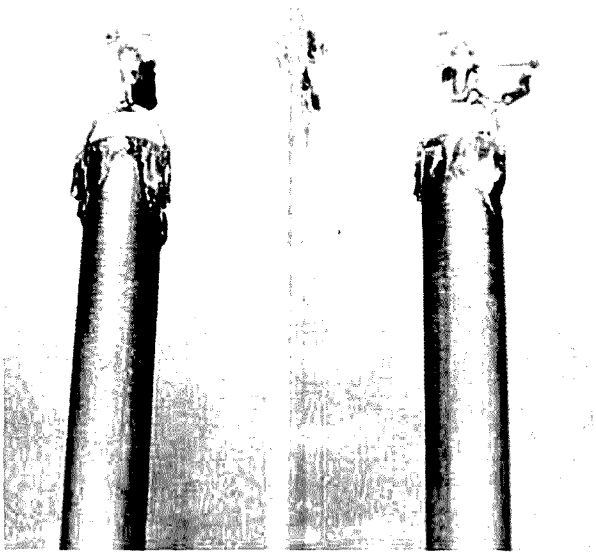
With the top plate removed from the chassis you can apply voltage to the unit. A meter in series with either the negative or positive power supply lead will allow you to monitor the current drain of the 2N5397. Correct current drain is 5 mA.* If you have more than this with a 330-ohm resistor at R1, increase R1 to the next standard value. If you read less than 5 mA lower the value of R1 (rule of thumb only—variables between 2N5397 transistors can reverse this suggestion). Keep adjusting the value of R1 until current drain is as close to 5 mA as possible.

To tune or test the unit the top plate

*Although this discussion is based on the 2N5397 fet, the less expensive 2N4416 can also be used in this preamplifier. However, the 2N4416 will have considerably higher noise figure, particularly above 100 MHz, and slightly lower gain than the 2N5397. The only change concerns the value of R1—it should be adjusted so that the current drain of the *2N4416* does not exceed 2.5 to 3 mA. The 2N4416 base diagram is the same as the 2N5397.

must be installed firmly in position with all of the sheet-metal screws. Connect a signal source (50 to 100 $\mu$V) to the input and a receiver or converter to the output. Slowly rotate C5 (stripline 4) for an indication of signal. At 50 MHz C5 will be about two-thirds meshed, on 144 MHz it will be half meshed and on 220 MHz it will be nearly wide open. If you can't find the signal (be sure your receiver is on the correct frequency) tune C3 (stripline 2) to the same position suggested for C5.

Once you find the signal, tune C2 and



**The cold ends of tubular striplines 3 and 4 are soldered to FT-1000 feedthrough capacitors.**

C7, then C6 and C4, in that order, for maximum signal. Peak the capacitors again, in the same sequence, starting with C5. (Crank down the output of the signal generator so you keep the S-meter in a more linear range.) With the unit tuned up for maximum gain, you will have an rf bandpass window similar to that shown in fig. 12. Fortunately this occurs at the same point as maximum gain.

Set the signal generator output so the S-meter reads S9. Disconnect the filter/preamplifier and connect the signal generator directly to the receiver. If the preamplifier is working properly—and the signal generator has a 50-ohm output—the S-meter should drop 10 to 14 dB, depending on the band you're .working with.

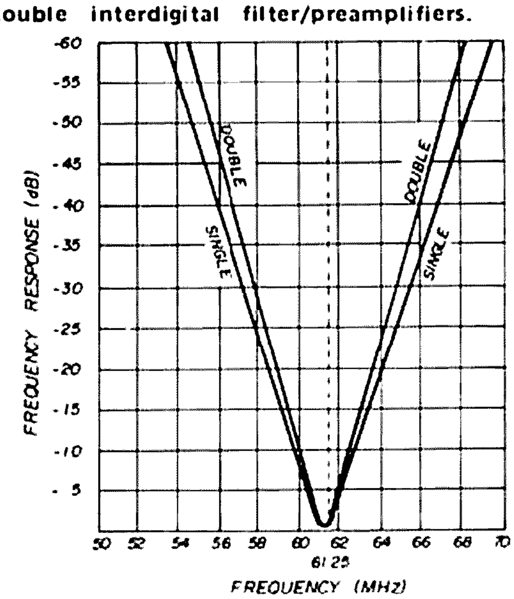
## precautions

Although I haven't had any trouble with high-power rf from my transmitter, you don't want to blow out the $8.50 transistor. Regardless of how you handle the potential problem—diodes, coax relay, etc.—do it *outside* the filter/preamplifier chassis. Any additional stray parts mounted inside the chassis only degrade performance by upsetting the balance of the tuned circuits inside the box.

## typical performance

Fig. 13A shows how a 10,000 $\mu$V channel-4 television signal appeared on a tv receiver tuned to channel 3. All of the video information on the screen is from channel 4; overload from channel 4 to channel 3 is virtually complete. Fig. 13B shows the same receiver, still tuned to channel 3, with an interdigital-preamplifier ahead of the receiver. The weak snowy picture, all 0.9 microvolts of it, is from a channel-3 transmitter 153 miles away. The 10,000-$\mu$V channel-4 signal has been completely eliminated. Fig. 13C shows the same signal passed through a pair of channel-3 interdigital-preamplifiers. With such a weak signal to work with there's not much hope for a snow-free picture. The signal input to the receiver at this point is 32 $\mu$V. (In this photo a 20-dB post amplifier was installed between the tv set and the second

**fig. 12. Bandpass curves for single and double interdigital filter/preamplifiers.**



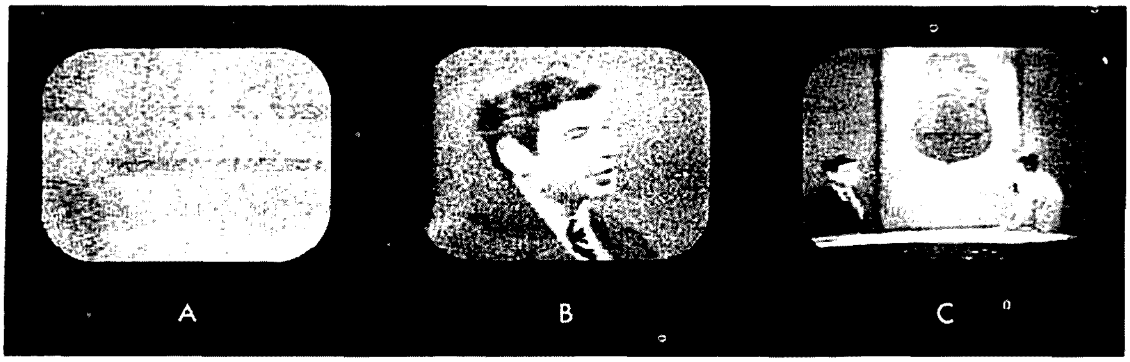


fig. 13. Adjacent-channel performance of the interdigital filter/preamplifier. In A the channel-3 programming is completely obliterated by channel-4 signal. With one interdigital filter/preamplifier in the line, channel-4 interference is eliminated (B). Two filter/preamplifiers improve the quality of the channel-3 signal (C).

interdigital-preamplifier.)

Fig. 14 shows the performance of the unit as a preamplifier without adjacent-channel interference. Fig. 14A shows an off-the-air signal from a channel-12 transmitter 170 miles away. If you look closely you may be able to see the frame bar. Fig. 14B shows the same signal with a single interdigital-preamplifier ahead of the receiver: now we know there's a signal present. Adding another interdigital-preamplifier results in the picture shown in fig. 14C.

It should be noted that the television receiver used for these photographs is a special catv model, the Conrac AV-12E. Comparing this receiver to the typical set you have in your home is like comparing a 75A4 to an S-38!

## sweep tests

The bandpass characteristics of the interdigital-preamplifier are shown in the oscilloscope displays of fig. 15. For these displays the output from a Jerrold 601D sweep generator, sweeping from 15 to 100 MHz, is connected through a 1-dB step attenuator to the input of an interdigital-preamplifier tuned to 61.25 MHz. The display is 10 dB from the baseline (top) to the bottom. The 62.25-MHz marker (1 MHz off resonance) on the right is 5 dB down the curve. Fig. 15B shows the same 62.25-MHz marker after passing through a pair of interdigital-preamplifiers; it is now 8 dB down the curve.

Fig. 15C shows the same curve with the marker moved down to 61.75 MHz, 500 kHz off the resonance point of the interdigital-preamplifier; it is 3 dB down the curve. The 61.75-MHz marker moves 6 dB down the curve when a second interdigital-preamplifier is added to the line (fig. 15D). From these measurements you can plot the bandpass curves of single

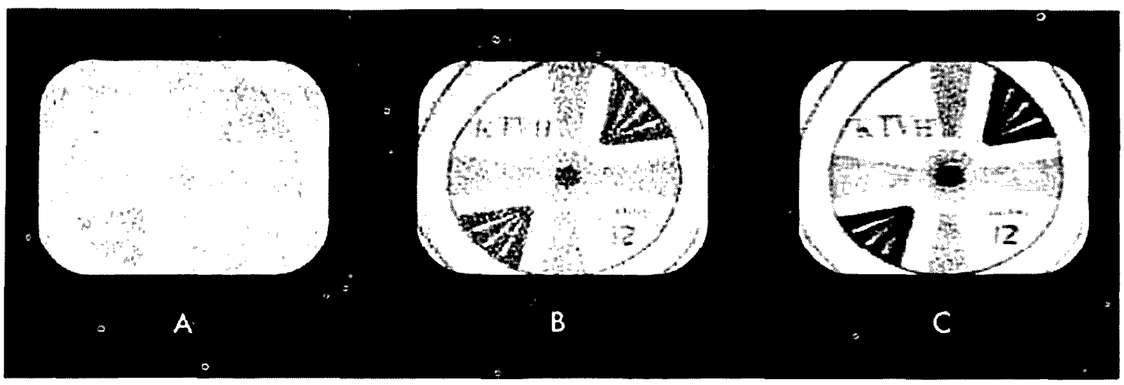
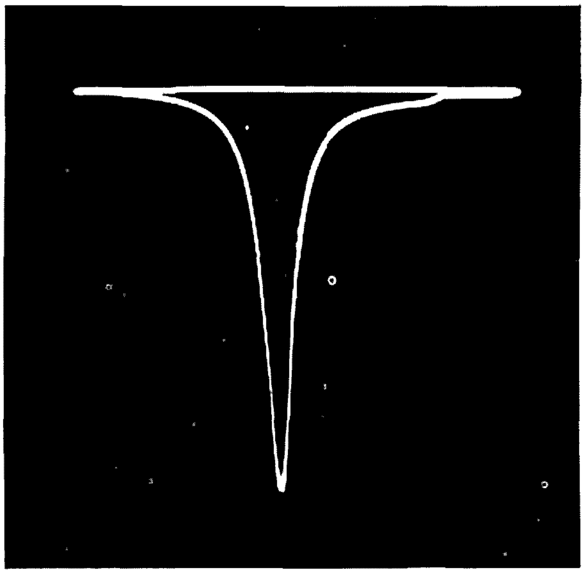
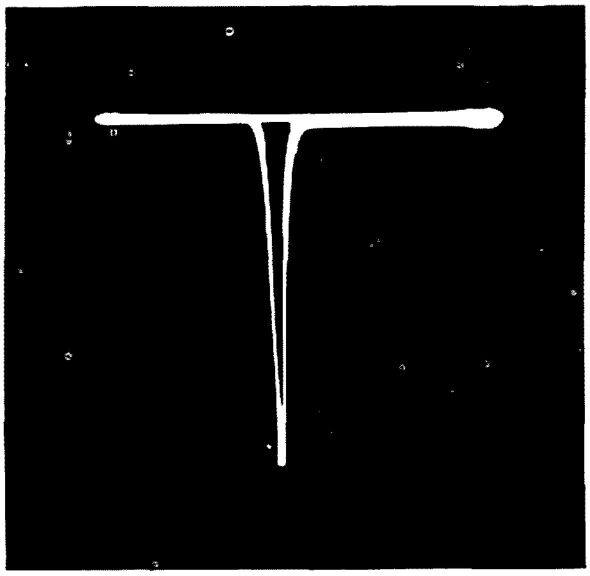


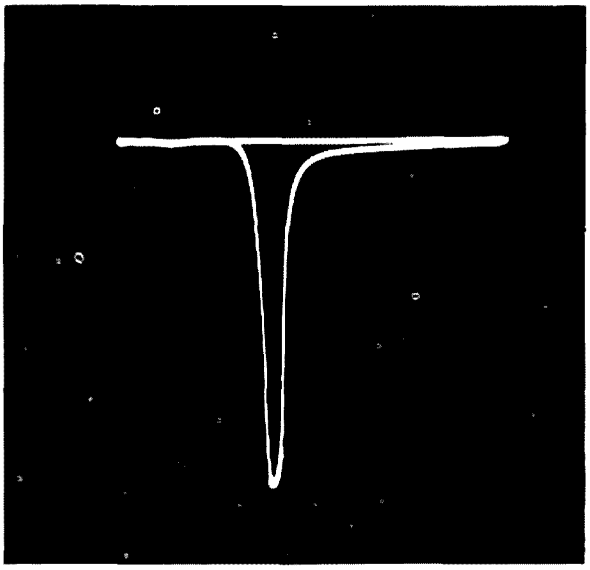
fig. 14. On-the-air performance of the filter/preamplifier. The channel-12 signal in A without the preamplifier in the line is almost covered with snow. One filter/preamplifier results in the signal shown in B. Second preamplifier ahead of the tv set results in the signal shown in C.



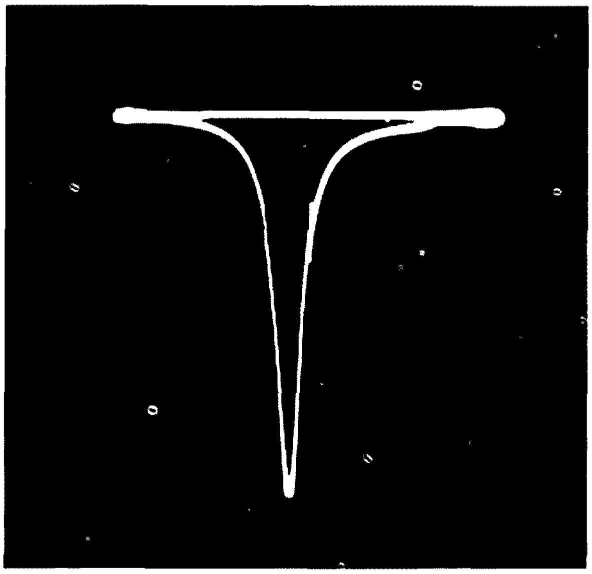
A. With single stage 1-MHz off-resonance marker is 5 dB down.



B. With two stages the 1-MHz marker is 8 dB down.



C. 500-kHz off-resonance marker is 3 dB down with 1 stage.



D. 500-kHz off-resonance marker is 6 dB down with 2 stages.

fig. 15. Bandpass characteristics of the filter/preamplifier.

and double interdigital-preamplifiers as shown in fig. 12.


## acknowledgements

I would like to acknowledge the cooperation I received from Charlie Williams of Siliconix, and Harold Cobbs of the W. Pat Fralia Company for their assistance in developing the 2N5397 parameters for this application. I am also indebted to Jay Liebman, W5ORH, who observed, "... state of the art is strange ... Just when I had my two-meter preamplifier and converter down to a microminiature package I have to build a new two-meter preamplifier that is bigger and takes more rack space than my two-meter kilowatt final!"


## references


1. Reed E. Fisher, W2CQH, "Interdigital Bandpass Filters for Amateur VHF/UHF Applications," *QST,* March, 1968, p. 32.
2. Reed E. Fisher, W2CQH, "Combline VHF Bandpass Filters," *QST,* December, 1968, p. 44.
3. Kenneth E. Holladay, K6HCP, "Super Two-Meter Preamp," *ham radio,* November, 1969, p. 72.
4. J. B. Compton, "High-Frequency Characterization and Application of Junction FETs," Siliconix Applications Note.


ham radio

# practical vxo design

An interesting approach to frequency stability in oscillator circuits

Gus Gerke, K6BIJ, Box 143, Weimar, California 95736

**You're on the air** having an enjoyable conversation. You switch over to the other station and the fellow says, "Sorry, missed most of that. Someone drifted onto your frequency." Sound familiar? The "someone" is usually a combination of unstable vfos and receiver drift.

The drifting signals one hears today suggest that vfo stability isn't really as good as claimed by equipment manufacturers and authors of vfo articles in the ham magazines. The best answer I've found to this problem is the variable-frequency crystal oscillator, or vxo.

The vxo circuits described in this article combine the flexibility (within limits) of a vfo with the inherent stability of crystal frequency control. Frequency can be varied between 2-720 kHz depending on the crystal frequency and other considerations, which I'll discuss. Many amateurs I've talked to never heard of varying a crystal's frequency over such a wide range.

Very little information has been written about the vxo. One article[1] describes a circuit that can pull down the frequency of an 8-MHz crystal about 4-5 kHz before the circuit becomes "a rather inferior vfo." With this circuit (**fig. 1**) as a starting point, I designed the circuits of **fig. 2** and **3** using FT-241 crystals in the 450-kHz region and the circuit of **fig. 4** using 3.5-8.5 MHz crystals.

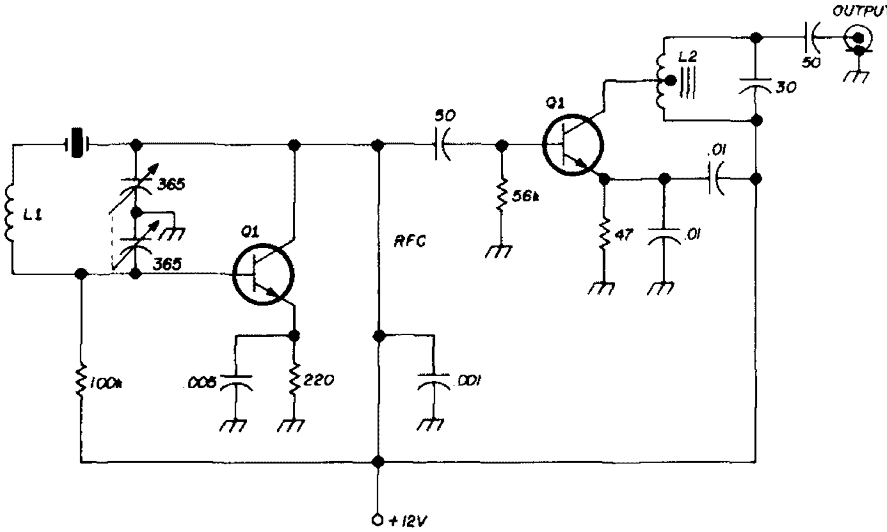
## circuit development

The vxo shown in **fig. 2** is a modification I made to a BC-604 fm tank transmitter. The vxo output goes through a stage of amplification and several frequency multipliers to obtain output on 21 MHz. I've used this vxo on 7 and 21 MHz cw with excellent results. The circuit has also been used to operate a 2-meter transmitter. Eight crystals were needed to cover the entire 2-meter band.

The only addition to the BC-604 was L1, C1. Capacitor C1 is used to pad the crystal frequency over a certain range, in this case 2 kHz. With an increase in padding range, the effects of temperature, vibration, and hand capacitance become

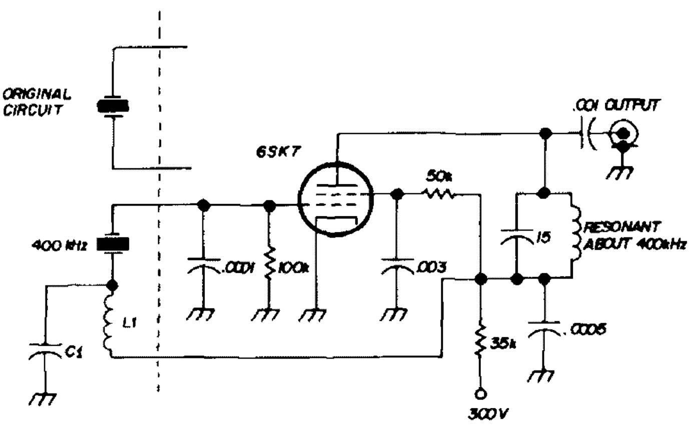
more pronounced; and the same precautions in building vfos must be used. These effects are small, however, and the crystal is still the frequency-controlling element. If you don't exceed the padding range, the vxo won't become an "inferior vfo."

The circuit of **fig. 3** seems to work well with the same low-frequency crystals used in the vxo of **fig. 2**. The solid-state version shown was also used with the BC-604. Since the crystals furnished with the BC-604 are less than 2 kHz apart, continuous coverage to the next lower-frequency crystal is possible. Stable 2-kHz padding was obtained with the circuit of **fig. 3**.




| | | | |
|---|---|---|---|
| L1 | 16-24 μH for 8-9 MHz crystal (J. W. Miller 4507) | Q1 | 2N706, 2N2219, 2N3662 or RCA 40237 |
| L2 | 40 turns no. 36, tapped at 16 turns | | |

**fig. 1. Circuit described in reference 1. An excursion of 4—5 kHz is claimed for an 8-MHz crystal.**




| | |
|---|---|
| C1 | Broadcast radio variable with both sections in parallel |
| L1 | Broadcast variloopstick antenna or similar |

**fig. 2. Oscillator modification made to a BC-604 transmitter using low-frequency crystal.**

A transistor vxo that produces stable 50-kHz padding is shown in **fig. 4**. This vxo can also be used with a crystal in the 8-MHz region for 6- or 2-meter operation. Doubling will produce a padding range of 100 kHz on 14 MHz, 150 kHz on 21 MHz with tripling, and 200 kHz on 28 MHz with quadrupling. To cover the entire 2-meter band, you'll need 8 crystals (500-kHz padding range).

**Table 1** gives recommended padding ranges for the FT-241 crystals when used in the circuits of **figs. 1** through **3**. If you're interested in a particular frequency range (as for net operation), try to use a crystal that will cover the first 25 percent of the padding range—then you'll have crystal stability.

The transistor circuits will start oscillating with 2.4 V; for more output, up to 12 V can be used. Unless followed by a frequency-multiplier, a buffer amplifier will be needed, as in **fig. 1**.

## a vxo for exciter use

Suppose you want to design a vxo

covering the entire 40-meter band, and you have an exciter such as the Central Electronics 20A using a 9-MHz crystal.

Higher than 9-MHz injection frequency is preferred to avoid unwanted mixer products. Therefore the injection frequency will be from 7 + 9 = 16 MHz to 7.3 + 9 = 16.3 MHz. Crystals in this range are overtone types and won't operate in these circuits. The solution is to use an 8.150-MHz crystal and operate it on its second harmonic, 16.3 MHz. Padding 50 kHz on the crystal fundamental frequency will produce 100-kHz shift in the output. This will give you full coverage of the 7-MHz phone band. An 8.1-MHz crystal will cover the next 100 kHz, and another crystal at 8.05 MHz will extend coverage to 7 MHz.

Crystals with frequencies of 8.125 and 8.075 MHz will be useful if you want extra stability and don't wish to pad more than 50 kHz on harmonics (25 kHz on the fundamental). These crystals are also useful for 2-meter work.

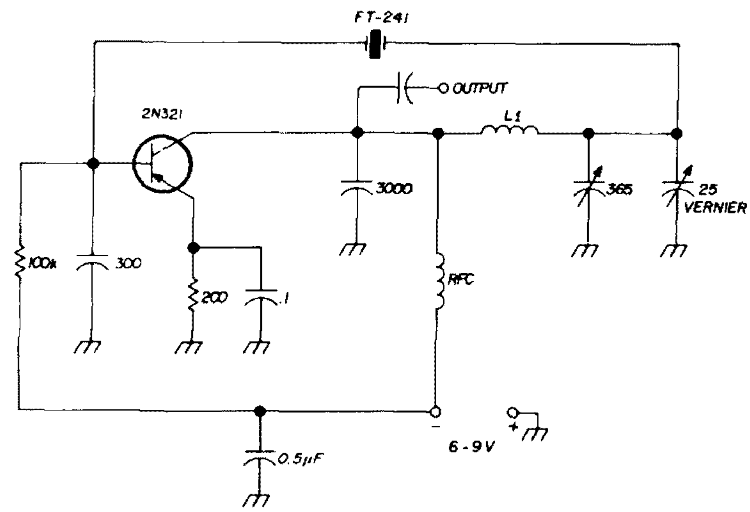



fig. 3. Solid-state version of the vxo in fig. 2.

## tuning capacitor considerations

Referring to fig. 5, capacitor C1 is used to bring the crystal frequency within the range of C2. Both capacitors should have a straight-line frequency response as a function of angle of rotation of the rotor plates. This capacitor characteristic is *important* for vxo calibration and tuning. For example, the tuning capacitors shown in the circuits of figs. 1 through 4 are common broadcast-band variables. When these are used, frequency decreases slowly at first as the capacitor rotor is turned. Then the frequency change becomes faster, until finally a hairline change in rotor position will produce a 1-kHz jump. This, of course, is very inconvenient at the lower frequen-

**L1** 40 turns no. 32 closewound on 3/8" slug-tuned coil form

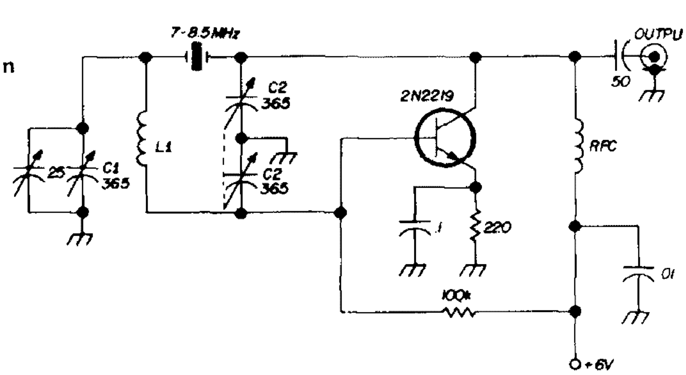



fig. 4. Solid-state vxo that produces stable 50-kHz padding on 7 MHz. It can be used for 6 or 2 meters also.

cies. The sketch of **fig. 6** illustrates the geometrical relationship of the stator plates in these two versions of variable capacitors.

In the circuit of **fig. 5**, capacitor C2 should be of good quality, otherwise contact-scraping noise will be heard in the receiver; small jumps in frequency may also occur. A capacitor with an insulated rotor is recommended for C2.

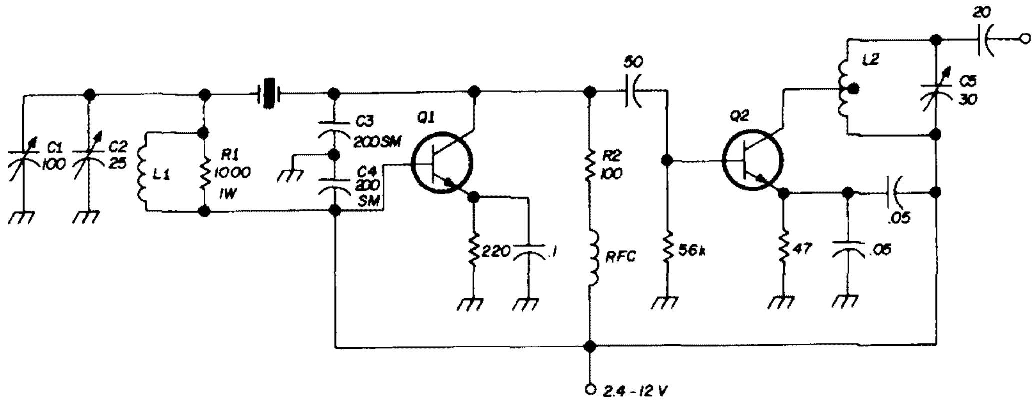



**fig. 5. Vxo-doubler circuit for a typical exciter. Three crystals are required for full coverage of the 40-meter band.**

## circuit description

The purpose of R1 in **fig. 5** is to lower the Q of L1. This allows a larger padding range and more stable operation near the low end of the range. If the frequency changes when touching the rf choke, the choke is too small. Resistor R2 prevents oscillation at the rf-choke resonant frequency.

Use a two-section bc variable capacitor to find the exact value of C3 and C4. Then replace the bc capacitor with two silver micas. A value of 200 pF seems right for this circuit.

**table 1. Padding ranges for various crystal frequencies.**

| FT-241 | padding range fig. 1 (kHz) | figs. 2 & 3 (kHz) |
|---|---|---|
| 0.45 (fundamental) | 0.2000 | 2.00 |
| 4.00 (9th harmonic) | 2.00 | 20.00 |
| 8.00 (18th harmonic) | 4.00 | 40.00 |
| 144.00 (324th harmonic) | 72.00 | 720.00 |

Battery voltage may be 2.4-12 V. Higher voltage may result in drift due to heating. I use 6 volts in my vxo.

## in closing

As far as I know, the vxo designs described in this article have never been published before. The circuit for the 20A exciter has been used on 40 and 15 meters in both the cw and ssb mode. All reports were crystal quality, and all operators asked for the circuit diagrams; so I've presented them here to share with others. My old vfo has since drifted into the junk box.

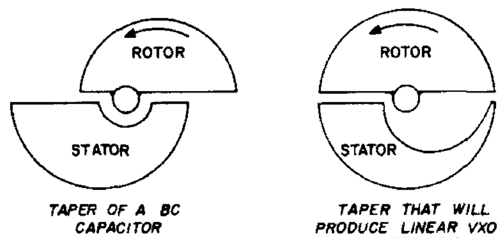
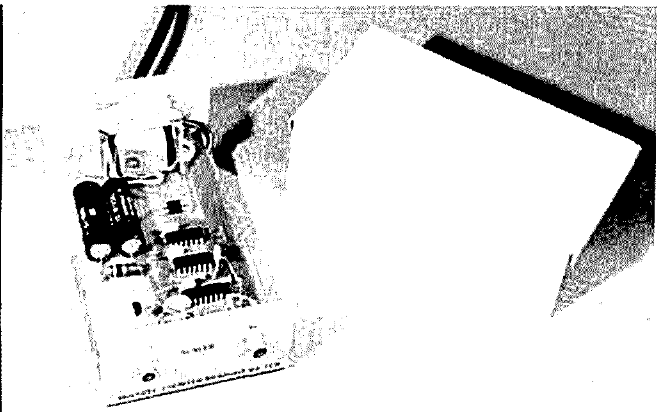



**fig. 6. Mechanical configuration of straight-line wavelength capacitor used for bc band and a straight-line frequency capacitor.**

### reference


1. J. R. Fisk, W1DTY, "73 Useful Transistor Circuits," *73*, March, 1967, p. 20A.


**ham radio**



# divide-by-ten frequency scaler

An accessory that will increase the range of your frequency counter by a factor of ten

Bert Kelley, K4EEU, 2307 South Clark Avenue, Tampa, Florida 33609

It's easy to extend the range of your frequency counter with this scaler. Four inexpensive JK flip-flops and a simple power supply comprise a circuit that will divide input frequency by ten. Thus, frequencies beyond the range of most home-built counters can be scaled down to a frequency the counter can handle. The basic measurement accuracy of the counter won't be affected by the scaler. If your counter has an upper measurement frequency limit of, say, 100 kHz, this scaler will extend its range to 1 MHz.

The maximum range of the scaler will depend on the IC's and construction techniques. The model shown here has an upper frequency limit of 106 MHz. Therefore, to realize the maximum potential of the scaler, your counter should operate to at least 10 MHz.

## applications

A frequency measuring system with a maximum upper range of 100 MHz will cover all amateur bands through six meters; however, the instrument isn't limited to these bands. Two-meter transmitters, for example, can be checked at a lower-frequency stage.

While exploring its possibilities, I used this scaler to check transmitters operating in the 900-MHz region by picking up rf at a frequency-multiplier stage. Transmitters operating between 144 and 160 MHz were also checked similarly. An insulated wire probe inserted near the plate coil was used to obtain energy for the scaler. Very little power is needed, and care should be taken not to overcouple the circuit. The IC's have a high-impedance input, and only 0.8 volt p-p is required to toggle them.

## circuit description

The scaler schematic (fig. 1) is a standard circuit for a clocked counter with binary-coded decimal output for decoding. The bcd output is incidental.

It's not used in the scaler, but merely means that the scaler is compatible with other IC's for decimal readout.

The IC's are from Motorola's MECL II family (meaning emitter-coupled logic). The MC1013P and MC1027P are identical, except the latter has lower internal such as the MC770P, for those interested in using an MECL as the first stage of a counter.

## construction

The main objectives in building the scaler were wide frequency response, sta-

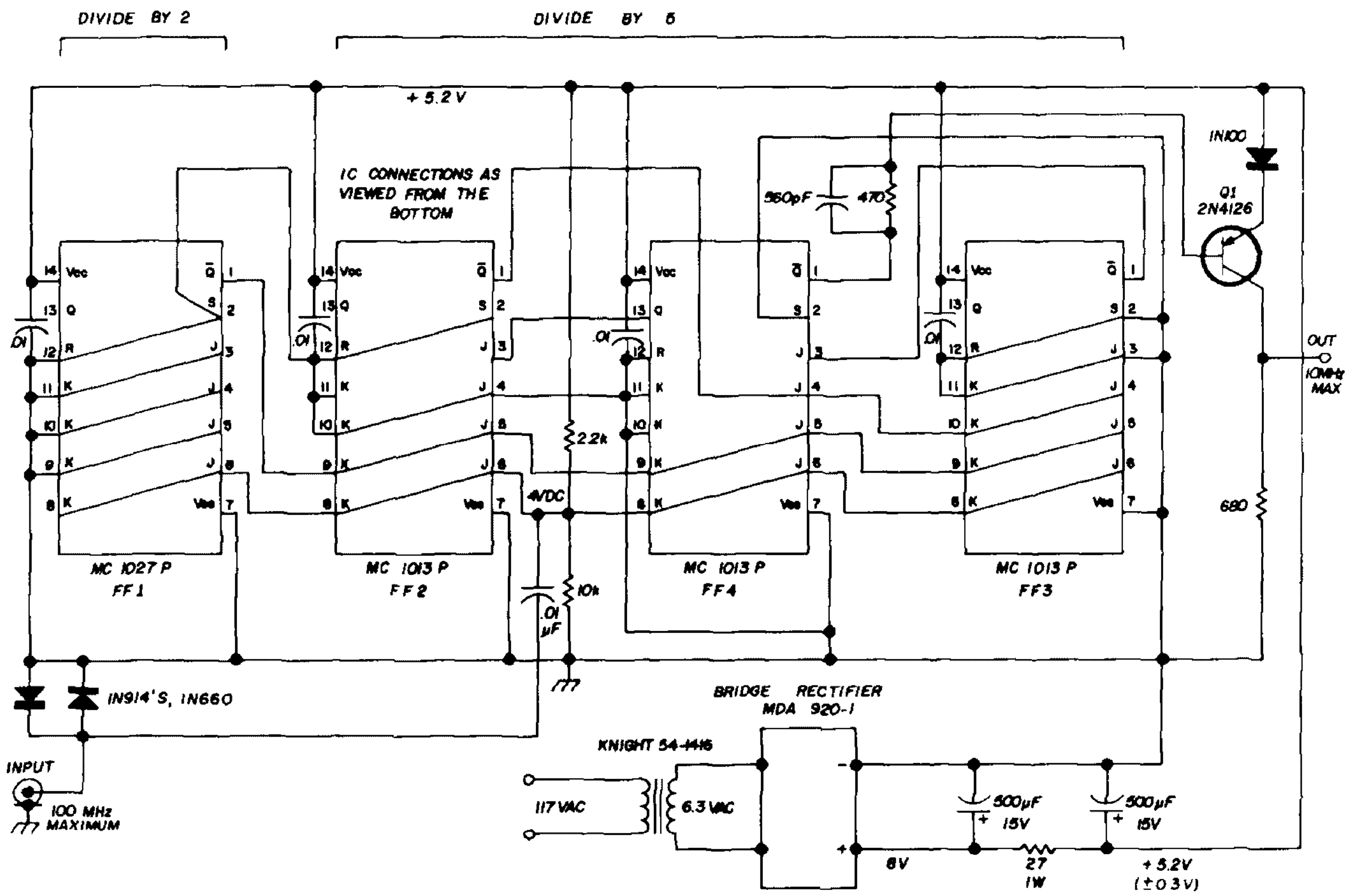



**fig. 1. Schematic of divide-by-ten scaler and power supply. Unit delivers 4 volts p-p to drive a digital frequency counter.**

resistance values, twice the current drain, and a higher operating frequency. I used the MC1027P in the divide-by-two section as a compromise between cost and performance. You can use all MC1013P's or all MC1027P's, since pin connections and logic are identical. Motorola technical data[1] may be consulted for details.

An amplifier stage, Q1, increases scaler output to 4 volts p-p. It will also translate the MECL logic level to saturated logic levels, so four identical circuits would be useful as an inexpensive interface between the bcd output and a resistor transistor logic (RTL) decimal decoder, bility, simplicity and low cost. I tried several other circuits using linear IC's, but found these JK's could be driven directly without gate drivers. The elimination of a preamplifier simplifies the power supply and reduces cost.

You are urged to use the board layout as shown in fig. 2, with no changes. Other layout arrangements will work, of course, but I arrived at this one after etching several circuit boards, and I know it will work.

At 100 MHz short leads or circuit trails are imperative, and bypassing is equally important. Note the 0.01-pF min-

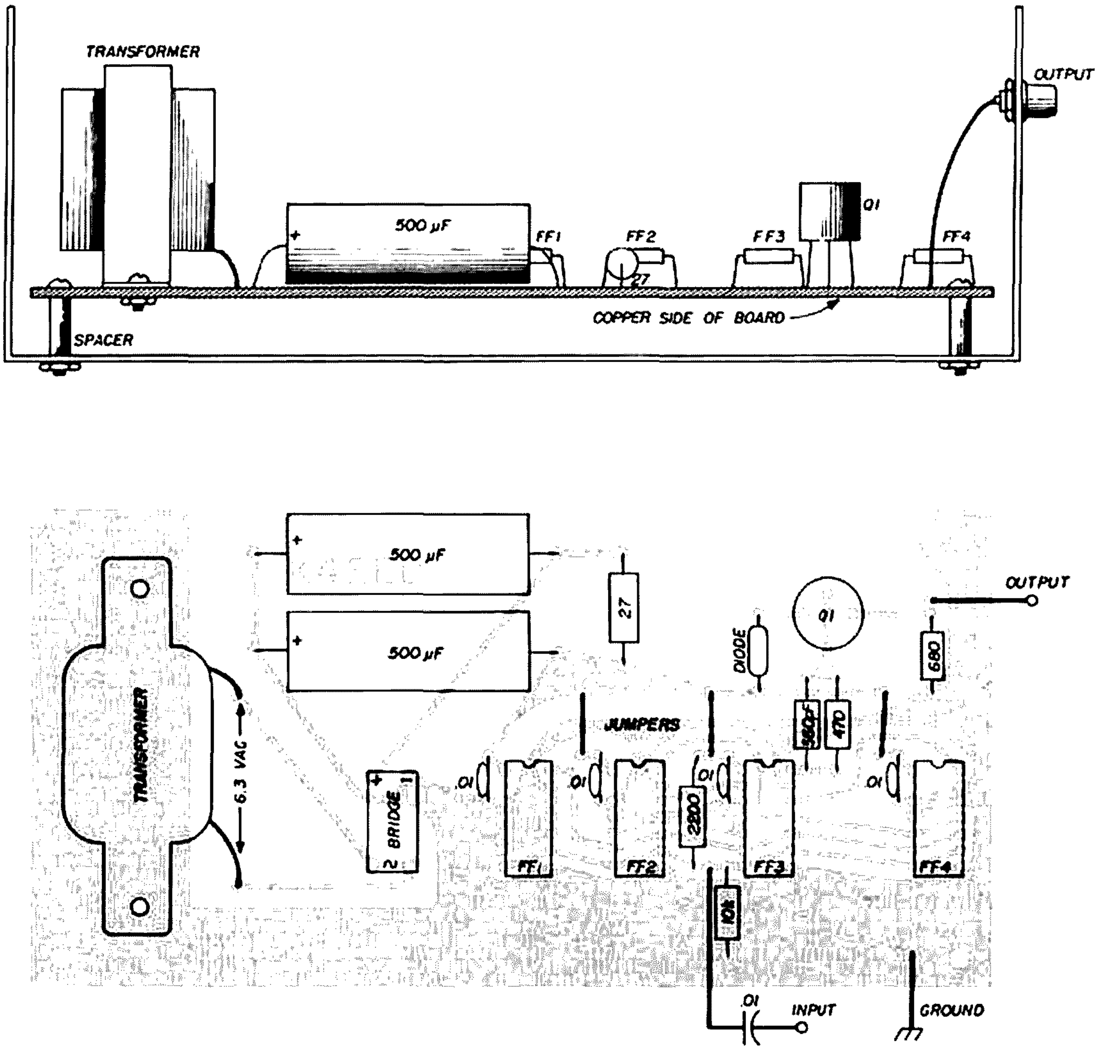



fig. 2. Parts layout. Enclosure is an LMB 138 box 6¼ x 3½ x 2-1/8 inches. Components are shown as viewed from the bottom of the board. Circuit boards are available from Stafford Electronics, 427 South Benbow Road, Greensboro, North Carolina 27401; order HR-7, $2.50.

iature discaps across the $V_{cc}$ terminals of each IC. Also it's important that power-supply output be close to 5.2 volts. The voltage at the junction of the 2.2k and 10k resistors must be 4.0 volts. Change the value of the 27-ohm filter resistor or the 2.2k divider until you obtain these voltages within 10%.

## wiring

When wiring the IC's be sure they're properly oriented and solder is flowed around each pin. Use a magnifying glass to examine each soldered connection to make certain that no pins are shorted. Use a 22-watt iron on the connection only long enough to ensure a good joint. A hotter iron can be used if you use some kind of device to transfer excess heat from the connection. Spring-loaded heat-sink tools are available for this purpose, or you can make an acceptable substitute from a strip of aluminum. Most device manufacturers provide maximum pin temperatures, and it's well to heed their recommendations.

If, for some reason, you must remove an IC you'll need something to remove the molten solder before the IC can be removed without damage. The best way is to use a vacuum device, such as a Sol-dapult,* to pick up molten solder. I've used this device when removing hundreds of solid-state devices and have yet to damage any from heat.

A complete parts list is provided in table 1. A conservative estimate for the total cost is $25.00—not bad for a simple circuit that will increase the range of your counter by a factor of ten.

*Edsyn, Inc., 15954 Arminta Street, Van Nuys, California 91406.

## operation

When the scaler is completed, check the power supply and examine the scaler circuit carefully. The JK flip-flops seem to operate with any input from 1 MHz up, so the scaler can be checked at a low frequency with a signal generator and inexpensive oscilloscope. If you obtain a signal at the output jack, the scaler is probably working all right. If not, it's easy to signal trace with the scope, progressing through flip-flops 1 through 4. Output should be found at pins 1 and 13 of each IC, at about 0.8 volt p-p. At the active pins (those not grounded or at $V_{CC}$), you should be able to measure 4.0 volts dc.

If scope checks lead you to suspect an IC, you can substitute another device to see if anything changes. With such a simple circuit, it's unlikely you'll have trouble; however the advice given above is included "just in case."

The two back-to-back diodes across the input provide some overload protection, but it's possible to zap all the IC's by applying a very high-level input signal. Therefore, with the scaler connected to the counter, increase scaler input until the counter suddenly starts reading; that's all the input signal necessary. A one-turn loop at the end of a short length of shielded cable makes a good coupling device.

**table 1. Parts list.**

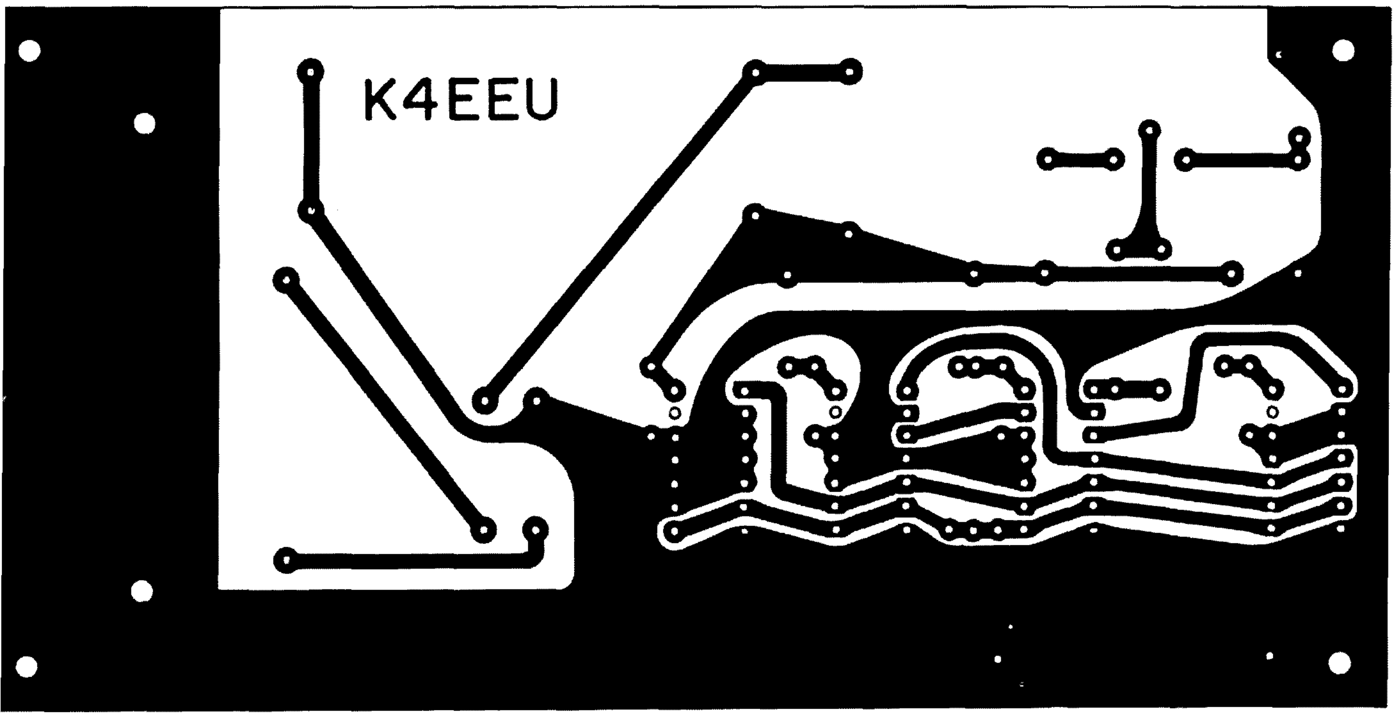
| | |
|---|---|
| 1 | 3 x 6-inch printed circuit board, copper one side, etched |
| 1 | enclosure, LMB 138 6¼ x 3½ x 2-1/8 inches or equivalent |
| 4 | spacers, 3/8 to ½ inch long |
| 1 | power transformer, Knight 54-1416 6.3V @ .6A (Allied Electronics) |
| 1 | single-phase bridge, or 4 diode equivalent (Motorola) MDA920-1 |
| 2 | 500 µF @ 15V (CD 500/15) |
| 1 | 27 ohm 1 watt resistor |
| 1 | 2.2k ½ W resistor |
| 1 | 10k ½ W resistor |
| 1 | 680 ohm ½ W resistor |
| 1 | 470 ohm ½ W resistor |
| 1 | 560 pF disc capacitor |
| 5 | 0.01 µF 100V disc ceramic or equivalent |
| 2 | 1N914 high speed diodes |
| 1 | 1N100 diode |
| 1 | 2N4126 pnp silicon transistor |
| 1 | integrated circuit (Motorola MC1027P) |
| 3 | integrated circuits (Motorola MC1013P) |
| 2 | phono chassis connector jacks |

### reference

1. "The Integrated Circuit Handbook," Motorola Semiconductor Products, Inc., P.O. Box 20912, Phoenix, Arizona 85306
Also, Application Notes AN257, AN224, and AN277.

ham radio

Circuit board for 10:1 frequency scaler.



# computer-aided circuit analysis

A powerful tool that eliminates trial and error in circuit design

M. A. Ellis, K1ORV, 61 Marlboro Road, Sudbury, Massachusetts 01776

**The number of unusual tasks** that can be more accurately and economically performed by a computer than by manual methods is steadily increasing. One of the more useful applications of the computer is the automatic analysis of electronic circuits.

A number of different circuit analysis programs are available. One of the first and still popular programs, developed by IBM, is ECAP (Electronic Circuit Analysis Program).[1,2] This program is widely used by circuit designers and has been adapted for use with machines other than those made by IBM.

## features

ECAP can be used for dc, ac, or transient analysis by making slight changes to the input data. One of the more useful features is an automatic frequency response analysis, with machine plotting of the waveshape. This is done by inputting the proper data for an ac or transient analysis and asking for an output plot.

As you would expect, a simple dc analysis is less complex than an ac or transient analysis of the same circuit. The dc analysis program for ECAP provides the steady-state solution for linear circuits composed of resistors, fixed current sources, fixed voltages, and dependent current sources. All other circuit components must be replaced with the equivalent circuit formed by these elements. The program also prints out diagnostics to inform the user of input errors and gives suggested remedies.

## application

To see how the program works, let's make a dc analysis of a simple two-stage transistor amplifier. We start with the schematic shown in **fig. 1**. The first step is to convert the schematic into a dc equivalent circuit so we can describe the circuit in the language of the computer.

It's necessary to follow a few simple rules to draw the equivalent dc circuit correctly. Capacitors must be replaced

with an open circuit and inductors with a short circuit, since this is how they appear to dc. The new circuit will consist of a network of junctions (nodes) and branches, each of which will have an input and output terminal. We call a node any point where two or more branches meet. (They are generally referred to as voltage nodes, since voltage calculations are made for these points as part of the mathematical solution.) A branch must contain at least one passive element—in our case a resistor. The branch may also contain voltage or current sources, or a dependent current source.

Each node and branch must be numbered. Any order will do, but we must begin with number 1 and not skip any numbers. Ground will be called node zero ($V_0$). Current-flow direction must be selected and may be arbitrary except for dependent current sources (the transistor collector current) and their controlling branches (base current branch); then the direction of flow must be consistent.

For example, the dependent current source in the transistor model must show current flowing from emitter to collector ($h_{FE} I_B$) when controlling current ($I_B$) flows from emitter to base. Refer to **fig. 2**, which illustrates the equivalent circuit model for the transistors in our circuit.

**Fig. 3** shows the equivalent dc circuit, with the two transistors (T1,T2) replaced by their models. The capacitors and inductors have disappeared; and the voltage nodes, Vx, and current branches, Bx, have all been numbered.

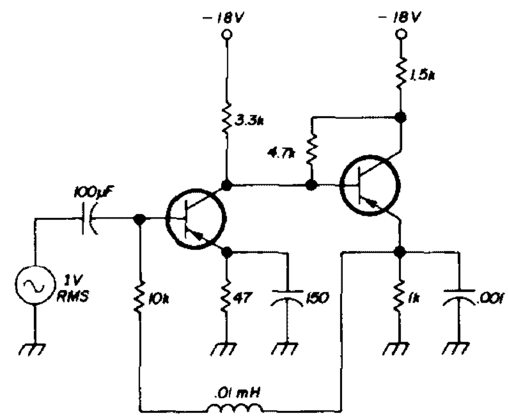



**fig. 1. Two-stage amplifier used in the circuit analysis. Transistor gain (beta) = 50, output impedance = 50k, input impedance = 200 ohms.**

Many readers will wonder why an article of this type is presented, since most amateurs just don't happen to have a digital computer handy. The fact is that the program described here is available through numerous software companies, who maintain remote terminals in many large cities linked by telephone line to a central computer facility. The computer is used by these companies and others on a time-sharing basis; thus cost to a terminal subscriber is prorated. Libraries of programs for many scientific and business-oriented problems are available.

The analysis program described here was developed for the IBM 1620 computer about eight years ago. It's only recently, however, that the cost of using the program has been within the means of the average user by time sharing. Typically, the cost of running this program would be about $15.00 an hour, including terminal occupancy and computing time. Within ten years or so this will be substantially lower, so you could have a time-share terminal in your home for perhaps the same price as a telephone subscription. **editor.**

## program analysis

We now start writing our input listing as pictured in **table 1**. We bypass the usual preliminary instructions to the computer concerning items not important here. "DC" (first line) indicates a steady-state dc analysis request. The second line tells the computer that in branch 1, current flows from node 2 to node 1 through a resistance of 200 ohms, and the resistor is connected to a battery voltage of -0.5 volt. (This represents the base-to-emitter voltage.) The third line describes current flow in branch 2 from node 0 (ground) to node 2 (emitter terminal) through a resistance of 47 ohms. Line four shows that, in branch 3, current flows in the dependent-current source from node 2 (the emitter-collector junction) to node 3 (the collector terminal), through a resistance of 50k ohms. E3 in computer language means multiply by $10^3$; i.e. three decimal places

table 1. Computer input listing for the electronic circuit analysis program.

| | |
|---|---|
| | DC |
| B1 | N(2,1),R=200,E=-0.5 |
| B2 | N(0,2),R=47 |
| B3 | N(2,3),R=50E3 |
| B4 | N(3,0),R=3.3E3,E=18 |
| B5 | N(1,4),R=10E3 |
| B6 | N(4,3),R=200,E=-0.5 |
| B7 | N(5,3),R=4.7E3 |
| B8 | N(0,4),R=1E3 |
| B9 | N(4,5),R=50E3 |
| B10 | N(5,0),R=1.5E3,E=18 |
| T1 | B(1,3),BETA=50 |
| T2 | B(6,9),BETA=50 |
| | PRINT,NV,BP |
| | EXECUTE |

should be added to the number 50. Line five depicts current flow from node 3 to node 0 through a resistance of 3.3k ohms and a battery voltage of 18 volts.

If the battery polarities seem confusing, remember that current flow from negative to positive through the battery represents an *increase* in voltage. Therefore E = 18 is a positive voltage. The opposite occurs in branch 1 when emitter-base current represents a decrease in voltage; hence, the battery is shown with current flowing into the positive terminal and out the negative terminal. Consequently the battery is represented by E = -0.5 volt.

The remaining branch descriptions can be read down through the line beginning B10. Line T1 shows that the transistor currents flow through branches 1 and 3, and that the transistor current gain is 50. In this line, the controlling branch, B1, must be listed first followed by the dependent branch, B3. The line beginning T2 is read in the same way.

We now tell the computer that we want it to print out node voltages and branch power losses by the PRINT, NV, BP line. The word EXECUTE, which follows, tells it to perform the calculations and print out the results.

## output

The printout is shown in **table 2**. Node voltages are printed first, with the node numbers shown at the left. The voltages for nodes 1 through 4 are listed on the first line and that for node 5 on the second. Note that the E 01 listed after the value for node voltage 3 means that the decimal point is to be moved one place to the right, indicating -2.64+volts. For node 5, E 02 states that we should move the decimal two places to the right.

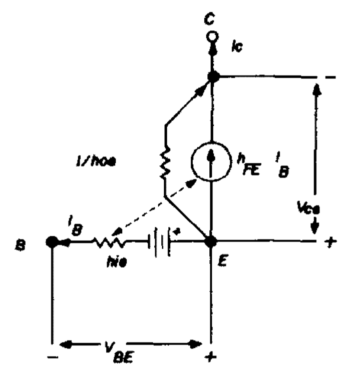



fig. 2. Equivalent circuit model for the transistors in fig. 1. The dashed arrow indicates the interdependence of the emitter-collector branch with the emitter-base branch.

table 2. Printout giving node voltages and branch power losses.

node voltages

| nodes | voltages | | | |
|---|---|---|---|---|
| 1- 4 | -0.83987243E-00 | -0.31386716E-00 | -0.26482825E-01 | -0.21401336E-01 |
| 5- 5 | -0.11750809E-02 | | | |

element power losses

| branches | power losses | | | |
|---|---|---|---|---|
| 1- 4 | 0.33813746E-05 | 0.20960125E-02 | 0.15285775E-01 | 0.71416734E-01 |
| 5- 8 | 0.16906791E-03 | 0.33202385E-06 | 0.17628932E-01 | 0.45801718E-02 |
| 9-10 | 0.21426438E-01 | 0.26034927E-01 | | |

Branch power losses are listed next in the same order as the node voltages. E-05 trailing the number for branch 1 tells us to move the decimal point 5 places to the left for a 0.00000338+ watt loss. Branch 2 shows a power loss of 0.002+ watt.

## modeling

You've probably guessed that the greatest difficulty with circuit analysis (or design) via the computer is coming up with realistic models for the active components, such as tubes and semiconductors. A good model allows useful results to be obtained, whereas a poor model will give questionable results or none at all. Almost as many different basic models are available as circuit analysis programs. Books have been written on modeling (references 3 through 6), therefore no attempt is made to go into that subject.

## other outputs

The circuit we've examined is simple. We have only asked for simple outputs. We could have asked for more profound information, such as voltage sensitivities or a worst-case analysis. Sensitivities refer to whatever change may occur in a node voltage for a one-percent change in the branch parameter. This allows the designer to take precautions and use closer tolerances with components that have the greatest effect on circuit operation. A worst-case analysis means to sum all the positive (or negative) tolerances in a parameter and compute the resulting node voltages. Obviously, these two outputs would be valuable to the circuit designer by eliminating many hours of manual calculations and the need for breadboarding.

## conclusions

Computer-aided circuit analysis or design is a powerful tool for the engineer or technician. With it he can design on paper, check out the circuit, and modify and recheck without breadboarding.

Those who have access to a computer may find it worthwhile to try their hand at circuit analysis with ECAP or one ot the other available programs. Anyone with a good understanding of electronic circuits can become proficient in computer-aided design and analysis techniques after a few practice sessions.

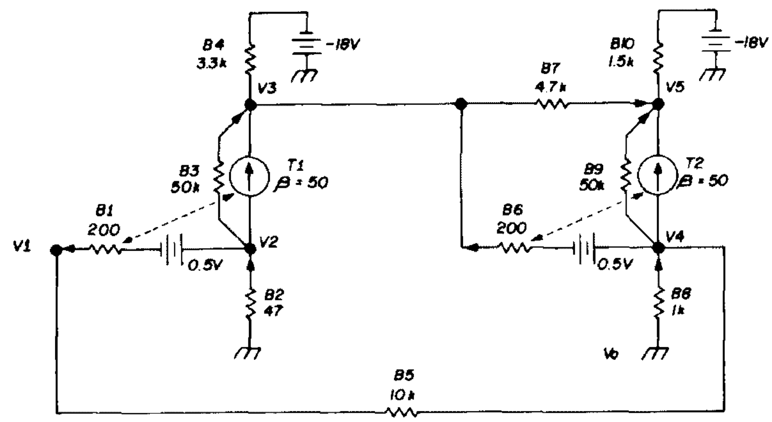



**fig. 3. Complete model of the two-stage amplifier with all the voltage nodes identified for the program input listing.**


### references

1. "IBM 1620 Electronic Circuit Analysis Program (ECAP)," document 1620-EE-02X, User's Manual, IBM, Inc., New York.
2. R. W. Jensen and M. D. Lieberman, "IBM Electronic Circuit Analysis Program," Prentice-Hall, New York.
3. J. F. Pierce, "Semiconductor Junction Devices," Charles E. Merrill Books, Inc., Columbus, Ohio.
4. "Transistor Manual," 7th edition, 1964, General Electric Co., pp. 52, 53; 69-77.
5. F. F. Kuo and W. G. Magnuson, Jr., "Computer Oriented Circuit Design," Prentice-Hall, New York.
6. "The Semiconductor Data Book," 4th edition, 1969, Motorola Semiconductor Products, Inc., pp. AN-3 to AN-15.


**ham radio**

# a tunable audio filter for cw

N. J. Nicosia, WA1JSM, 85 North Street, North Reading, Massachusetts, 01864

**A selective audio filter can** really help to pull cw signals out of heavy interference. What will help more is a tunable unit that will allow you to select the tone you want.

This circuit uses two cascaded Raytheon RM709 linear operational amplifiers* in an active filter that can be tuned while still maintaining essentially constant bandwidth at the 3-dB points of its response curve.[1]

## design

**Fig. 1** shows the filter response at 1000 Hz using the circuit of **fig. 2**. Gain at center frequency is approximately zero dB (gain of one), and the tuning range is from 750 to about 1600 Hz. The 3-dB bandwidth is 140 Hz.

The gain of one means that when the filter is switched in, audio blasting won't occur at the tuned frequency. Maximum input signal is about 5 volts before clipping occurs at the output.

## power supply

The ICs require a dual-polarity power supply. My supply makes use of the trick known as "zero" or "common" reference to the IC.[2] A single 9-volt transistor battery with two resistors is used, as shown in **fig. 3**. While this supply is adequate, slightly higher voltages will allow the filter to handle larger input signals.

## construction

Careful parts layout and the usual construction practices for assembling and wiring integrated circuits are a must for this filter. The ICs have high gain and wide bandwidth. Short leads and bypassing at IC terminals will ensure against internal oscillation, which could destroy the devices. Overall shielding, as well as input-lead shielding, will keep transmitter rf out of the filter. The parts can be mounted on a small perforated board, which can then be installed in a Minibox or similar enclosure.

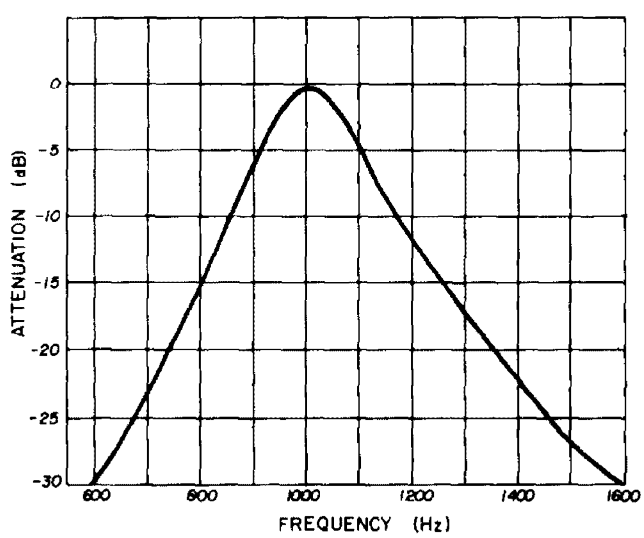
## operation

After checking filter and power supply wiring, set the filter switch to OUT, insert the filter phone plug into your receiver phone jack, and connect your headphones to the filter output. Tune in a cw signal tone of your choice, then switch the filter to IN and adjust the filter tuning for maximum volume. Select a tone within the filter tuning range. With practice this will be easy.

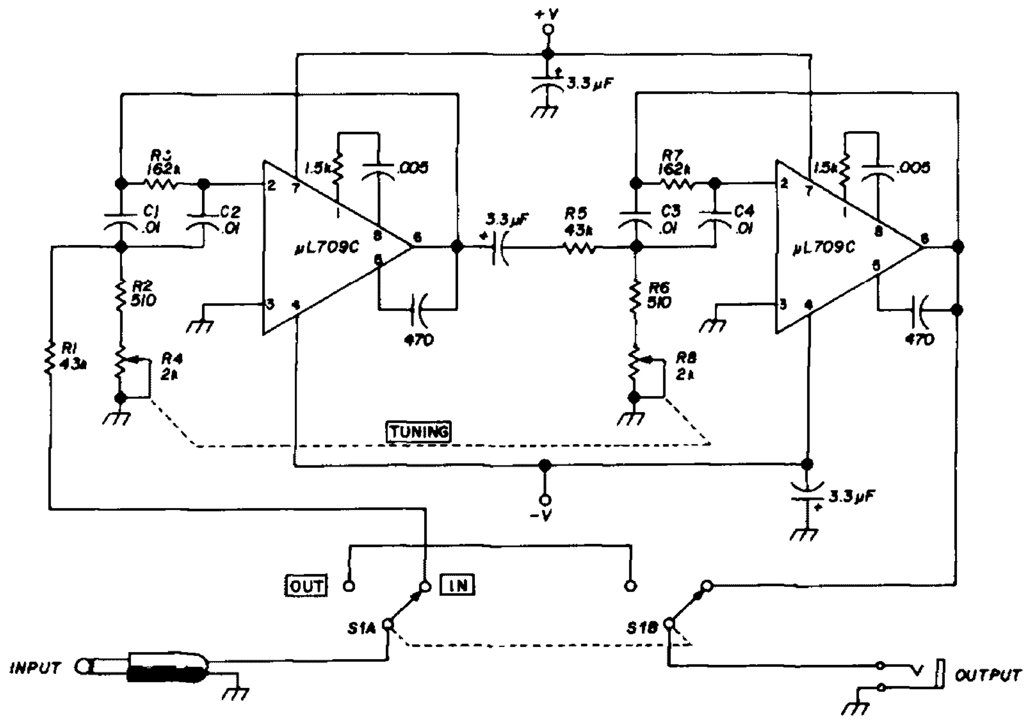
## performance

Although the filter skirts are not the best I have seen, the unit performs remarkably well. If your ssb transceiver

**fig. 1. Filter response with a center frequency of 1000 Hz.**




*Directly replaceable with the Motorola MC1709. Both devices are packaged in a TO-99 case. **Editor.**




| | | | |
|---|---|---|---|
| C1, C2, C3, C4 | .01 µF, matched to 5% or better | R3, R7 | 162k, matched to 2% or better |
| R1, R5 | 43k, matched to 2% or better | R4, R8 | 2k ganged potentiometers (IRC no. PQ11-110 and M11-110) |
| R2, R6 | 510 ohms, matched to 2% or better | | |

**fig. 2. Schematic of the tunable audio filter. Circuit features selective audio tone for cw reception.**

has a fixed receiver bandwidth (approximately 2.5 kHz) you'll be able to receive cw with considerable ease with this filter. The filter tuning control literally scans the receiver passband and picks out the wanted tone. Using the filter with my Swan 500C has made many enjoyable cw sessions possible.

## possibilities

Because the filter can handle high-level signals, it's not restricted to headphone use. With some thought it could be used for speaker operation.

For higher gain, a different audio range, or change in bandwidth, the component values can be adjusted in accordance with the following formulas:

$$Ra = \frac{1}{2\pi\Delta GC}$$

$$Rb = \frac{1}{\pi\Delta C}$$

$$Rc = \frac{\quad}{2\pi C\,(2f^2/\Delta) - \Delta G}$$

where

$\Delta$ = desired 3-dB bandwidth per stage
G = nominal stage gain at center frequency*
C = C1, C2, C3, C4
f = midrange frequency
Ra = R1, R5
Rb = R3, R7
Rc = R2 + R4; R6 + R8

*Total filter gain = gain product of stage 1 and stage 2.

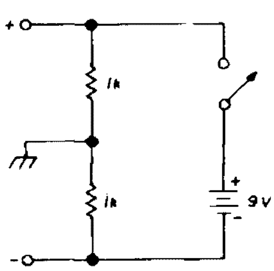
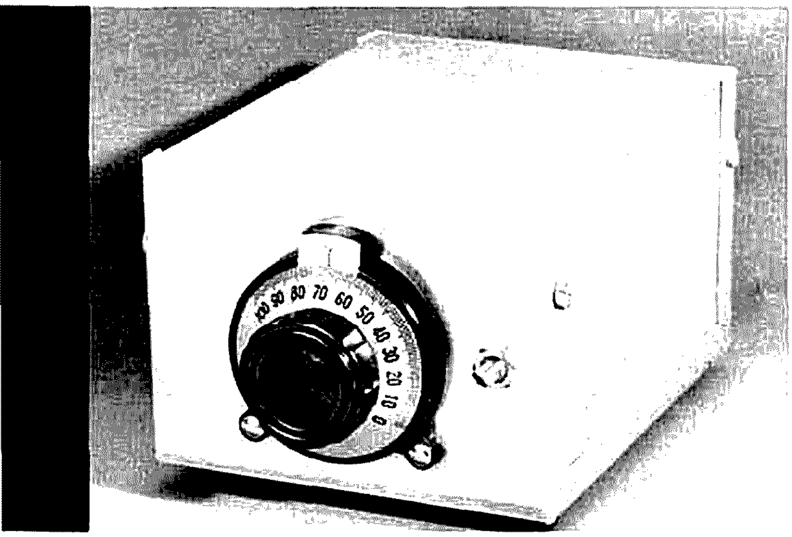



**fig. 3. Transistor-battery power supply that provides dual-polarity output for the filter.**

## references


1. Leslie Robinson, "Active Bandpass Filter with Adjustable Center Frequency and Constant Bandwidth," *EEE*, February, 1968, p. 124.
2. John J. Schultz, W2EEY, "Using Integrated circuits with Single-Polarity Power Supplies," *ham radio*, September, 1969, p. 35.


ham radio



# a vfo for solid-state transmitters

Tired of being rockbound? Here's a neat vfo featuring the Vackar oscillator

C. E. Galbreath, W3QBO, 8326 Still Spring Court, Bethesda, Maryland

Amateurs who have built and used very low power solid-state transmitters have marvelled at their ability to carry on solid contacts over normal propagation distances, even on 80 and 40 meters. Most of these transistor transmitters, however, are designed around crystal-controlled oscillators. The limitation of operating within the frequencies of one's supply of crystals is not relished by most amateurs accustomed to moving freely over the bands. In this age of the vfo, crystal control of operating frequency represents a step backward in amateur communications practice.

Having been bitten by the transistor transmitter bug but not wanting to be crystal bound, I began searching the literature for solid-state vfo construction articles. Fortunately there are a number of helpful articles on various transistorized vfo circuits (see espicially reference 1), and I will not discuss these circuits here. Unfortunately a dearth of *construction* articles exists for the amateur wanting to build a stable, easy-to-

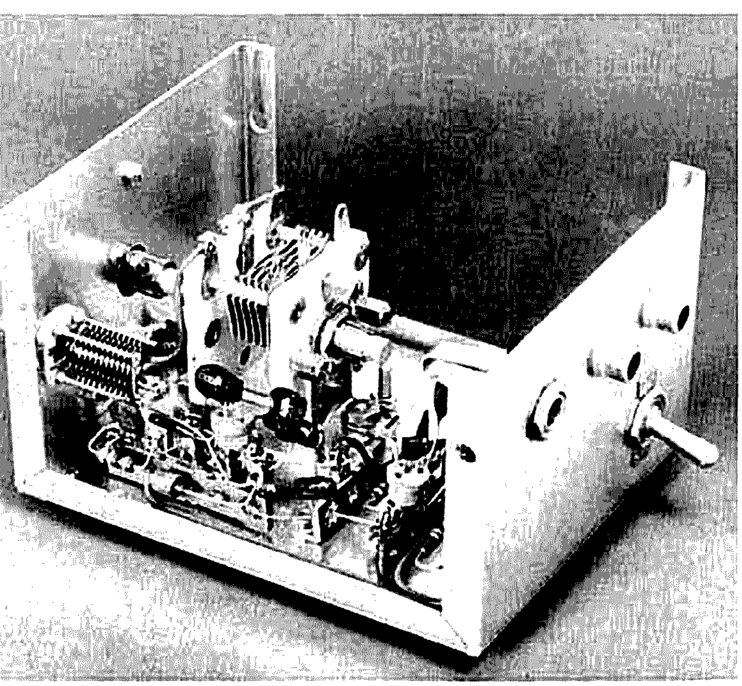
construct solid-state vfo. Most amateurs do not have the time or technical knowledge to design, work out the construction details, and conduct the tests necessary to build such a unit. These roadblocks limit the interest of many amateurs in transistor cw work. This article will help to fill a gap in the literature and provide an interesting construction project.

The vfo is easy to reproduce, produces a high-quality chirpless note when keyed, and is relatively inexpensive to build, even when all parts are purchased new.

Two vfos were constructed, one for 80 meters and one for 40. The 80-meter unit, however, can be used on 40 if a doubler stage is used in the transmitter. For those wishing to construct a 40-meter unit, appropriate oscillator component values are given. All other construction details are the same for both. In fact, the circuit can be used from 10 through 160 meters with appropriate changes in values of the frequency-controlling components.

The circuit is uncomplicated and easy to follow (fig. 1). It uses a Vackar oscillator with a toroidal inductor, a buffer stage for isolating the oscillator and minimizing the oscillator load, and a second buffer for further isolation as well as impedance matching. Inexpensive jfets, MPF102s, are used in the oscillator and in the first buffer stage, and a bipolar transistor, an HEP 55, is used in the impedance-matching stage.* The HEP 55, operating in the common-collector configuration, matches the high impedance output of the MPF102 buffer to the low-impedance input of the transmitter.

**Stable high-frequency vfo uses a double-bearing Millen variable capacitor.**



## construction

The vfo is constructed in a 3 x 4 x 5-inch aluminum minibox. Rubber mounting feet on the minibox help reduce mechanical shock and vibration. Also, a heavy aluminum plate is used to reinforce and strengthen the base of the minibox. The main tuning capacitor—a double bearing type—and the toroidal inductor are mounted on a 1½ x 2¾-inch platform of plexiglas or polystyrene supported on spacers on the base of the minibox. This method of mounting the capacitor and coil minimizes vibration and increases stability. It is an improvement over the conventional practice of direct mounting on a heavy metal base. As someone has pointed out, heavy metal, as in a bell, vibrates vigorously when even lightly tapped. This is not true of plexiglas or polystrene. The oscillator coil is wound on an Amidon T-50-2 toroid core and needs no shielding. It is mounted on a 1 x 1-inch piece of perforated board and is held firmly by coil dope and short leads soldered to lugs bolted to the board. The perforated board, in turn, is bolted to a small standoff insulator mounted on the plexiglas platform be-

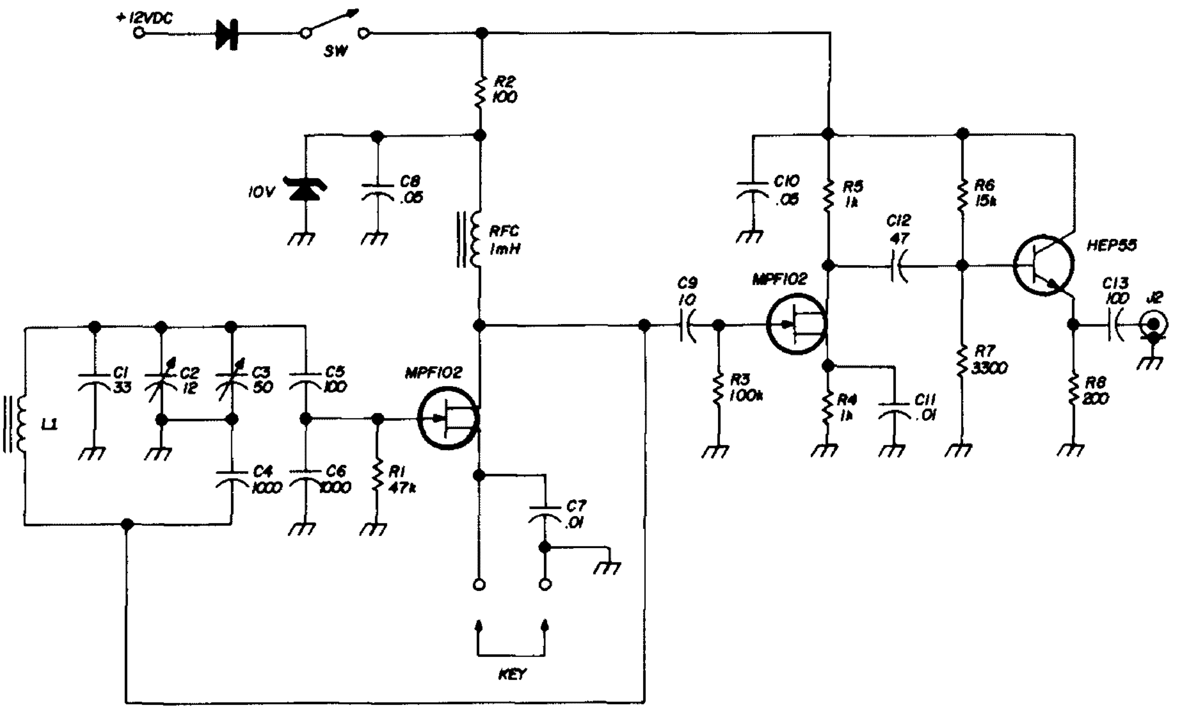
*The MPF102 is listed at 90 cents and HEP 55 at $1.20 in the current Allied Radio Company catalog.

tween the main tuning capacitor and the front of the minibox. The other oscillator components are mounted on terminal strips along one side of the plexiglas platform. A 10-volt zener provides excellent voltage regulation for the oscillator.

All components in each buffer stage are mounted on a single five-tie-point terminal strip, center terminal grounded. Each buffer stage is assembled as a unit before attaching to the chassis. Copper-clad circuit board is not recommended. Not only is it difficult and time-consuming to use, but it offers no performance or constructional advantage over conventional terminal-strip wiring in this application. In fact, circuit boards frequently cause ground-loops and other problems.[2]

Transistor sockets are used in preference to soldering the transistor leads into the circuit. This not only avoids possible damage to the transistors from the heat of the soldering iron, but also permits easy substitution of transistors if and when necessary. The latter feature paid off for me, as I found a weak transistor in the circuit when I first tested the vfo. In mounting each socket, two of its terminals are soldered to the appropriate tie points on the terminal strip.

The need for temperature compensation to reduce frequency drift of the oscillator is minimal so long as the vfo is kept away from heat-generating equipment. A 33 pF N750 negative-temperature-coefficient capacitor is sufficient to take care of small temperature-induced




**L1** (80 meters) 48 turns no. 30 enameled on ½" toroid core (Amidon T-50-2)

(40 meters) 25 turns no. 30 enameled on ½" toroid core (Amidon T-50-2)

**C1** 33 pF N750 negative temperature coefficient ceramic

**C3** (80 meters) Millen 23050MKF or 19050

(40 meters) Millen 19025

**RFC** Millen J300-1000

For 40-meter operation C4 and C6 should be 680 pF

**fig. 1. Schematic for solid-state vfo and buffer stages.**

effects.

Note that a subminiature iron-core choke is used to provide the load for the drain of the oscillator transistor. The choke replaces a resistor that was first tried. I could not get the circuit to oscillate on a 12-volt supply using the load resistor because of the voltage drop. The low value of the inductance (1 mH) of the choke not only provides low dc resistance but also the proper load impedance for the MPF102.

For adjusting the main tuning capacitor, an inexpensive 2-inch Japanese-made vernier dial is used. This works very well. Band edge is adjusted by a 12 pF variable capacitor on the front panel; a slotted shaft for screwdriver adjustment is recommended, as this capacitor rarely needs attention. The bandspread over the main tuning dial is excellent.

On the back of the minibox are the key jack, two phono jacks (one for rf output and one for battery or power supply leads) and an on-off switch for B+. To avoid damage to the transistors from reversed polarity, a semiconductor diode is inserted in series with the B+ lead between the B+ jack and the on-off switch.

The value of capacitor C12 coupling the vfo to the transmitter will depend on the amount of drive required for the transmitter and the input impedance of the crystal oscillator stage, whose crystal the vfo now replaces. Although the schematic shows a 100 pF coupling capacitor, the value might well vary from 50 to 300 pF.

## checking and testing

Before connecting the vfo to its power supply and before inserting the three transistors, use your vom to check for any possible short between the B+ line at the "on" side of the switch and the chassis. Still without inserting the transistors, connect the power supply to J1. Check with the vom for correct voltage between the "off" side of the switch and chassis. If the voltage is not correct, the diode is either inserted backwards or is defective. Next, insert an MPF102 in the oscillator-stage socket and connect a key to the key jack. Turn on the switch and, while keying, listen for the oscillator signal in your receiver. With vfo tuning dial set at 90 you should find the signal toward the low end of the cw portion of the 80-meter band; the exact spot will depend on the setting of C2.

When you have ascertained that the oscillator stage is functioning normally, turn off the switch and insert the other two transistors in their sockets. Also connect a 0–500 ohm or a 0–1000 ohm potentiometer or variable resistor as a load to the output at J2. Turn on the switch and with the key depressed measure the rectified output voltage, using a

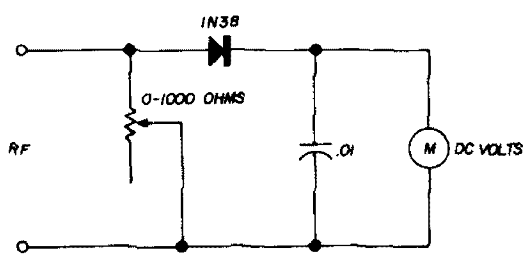



**fig. 2. Test probe for checking output of vfo.**

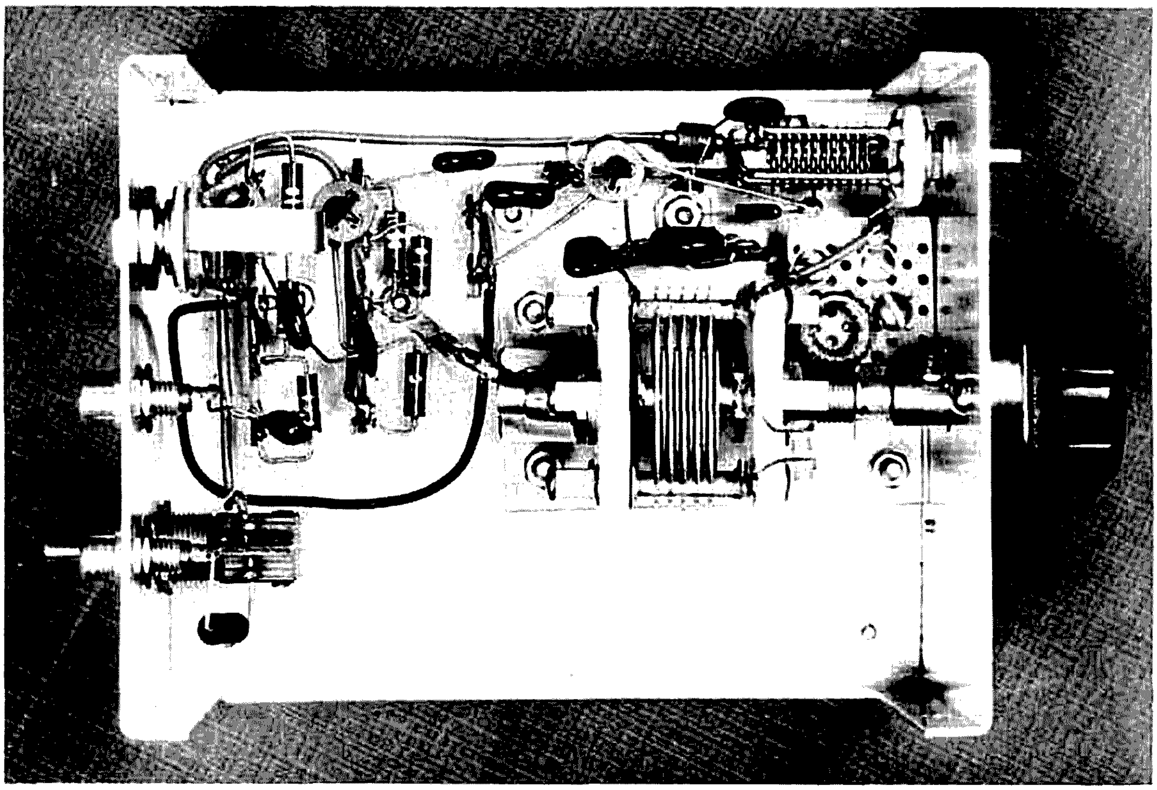
probe such as that in **fig. 2**. The output should peak about 1½ as the potentiometer is adjusted—probably around 400 ohms. One other voltage should also be checked: measure the regulated voltage at the zener; it should remain unchanged while keying.

When these tests have been successfully completed, you are ready to connect the vfo to the transmitter.

## connection to transmitter

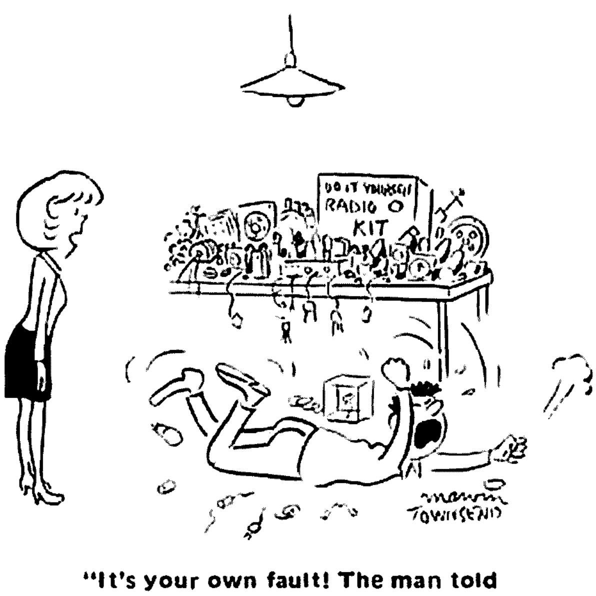
A short piece of coax or shielded wire can be used to connect the output of the vfo to the transmitter. The center of the coax is tied to the terminal of the crystal socket that is connected to the input of the oscillator (usually the base). Be certain that this connection is made correctly! Do not connect the other crystal socket terminal to the chassis unless it is already connected that way. The braid of the coax, of course, is tied to chassis

ground. It is recommended that a phono jack be installed in the transmitter for the connecting cable.



Parts layout for the stable solid-state vfo.

You may have to change coupling capacitor C13 to obtain the required voltage at the input of your transmitter.




"It's your own fault! The man told you this kit separates the men from the boys!"

The only other adjustment is to set the low-band edge on the main tuning dial. This is done by turning the main tuning dial to 100 (full capacity) and adjusting C2 while checking with the crystal in your receiver.

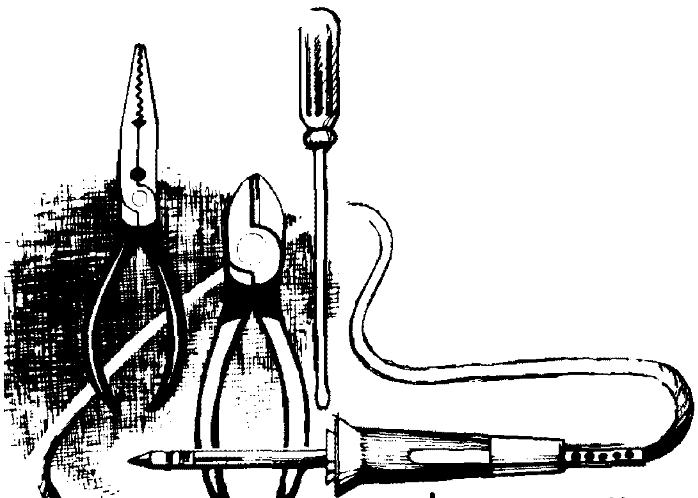
The vfo is powered by a 12-volt lantern battery or eight D-type flashlight cells. Be sure to connect a 100-$\mu$F electrolytic capacitor across the battery terminals.

I'd be happy to hear from those who build and use this vfo.

### references


1. J. R. Fisk, W1DTY, "Stable Transistor VFOs," *ham radio*, June, 1968, p. 14.
2. Jim Ashe, W1EZT, "Grounding and Wiring," *ham radio*, June, 1969, p. 44.
3. G. D. Hanchett, W2YM, "The Field-Effect Transistor as a Stable Vfo Element," *QST*, December, 1966, p. 11.
4. P. H. Lee, W3JHR, "A Stable Transistorized VFO," *CQ*, September, 1963, p. 25.
5. W. A. Wildenhein, W8YFB, "High-Stability Solid-State VFO," *ham radio*, March, 1969, p. 14.


ham radio



■ by larry allen

# the repair bench

## how distortion creeps in

"Whaaatsch at cherrr saoyynn naaiir, ohhhyemmm?" the speaker sputters. "Cauyynt unnderrrschtaunnnd moy kewwayysssaooh?"

Heck no, you can't understand his QSO. He sounds like his mouth's full of spaghetti. What's happening is the speech amplifier in that fine new high-power transmitter has a gremlin. He's being bugged by distortion.

When distortion hits you in your receiver audio amps, you at least have a sporting chance. You can hear it. But audio distortion in your transmitter is embarrassing. Everyone elso knows it before you do. (They do unless you're a lucky one who has a scope modulation monitor and watches it.) Whichever place it hits, the sooner you get at fixing it, the sooner you'll start enjoying your hobby again.

### the unlinear amp

Today's speech amps and audio amps are mostly transistor. A good percentage of the faults that happen pop up suddenly. You're working along and—bang, there's trouble.

Not so, when the trouble is distortion. Garbled sound has a way of slipping up on you. At first the sounds you (or your buddies on the net) hear are only vaguely "not right." But they get steadily worse over the next few hours or days—sometimes over weeks or months. Then one day you wake up to how hard it is to understand anybody on your once-hot receiver. Or people ask you to repeat so often it's annoying.

Preventive maintenance, particularly in the transmitter, may help some. If you're running ssb, it's awful easy to have distortion just from misadjustment. So it's a good idea to check out the speech amps whenever you feel the urge to make sure your signals are shipshape.

The chief cause of audio (or speech, or voice, or sound, whichever you like to call it) distortion is an amplifier that has become nonlinear. A class-A amp should operate on the linear portion of the base-collector (or grid-plate) transfer characteristic **(fig. 1)**. Let the bias shift, and operation moves onto one of the knees. The resulting distortion can be bad. See the sine-wave comparison in **fig. 2**.

Some speech drivers operate outside class A. They may run with bias nearer cutoff, like a class $AB_1$ **(fig. 1)**. Major distortion is avoided there by carefully limiting the amplitude of input signal and by using transformer output coupling. A transformer helps restore the roundness of a sine waveform that has been flat-

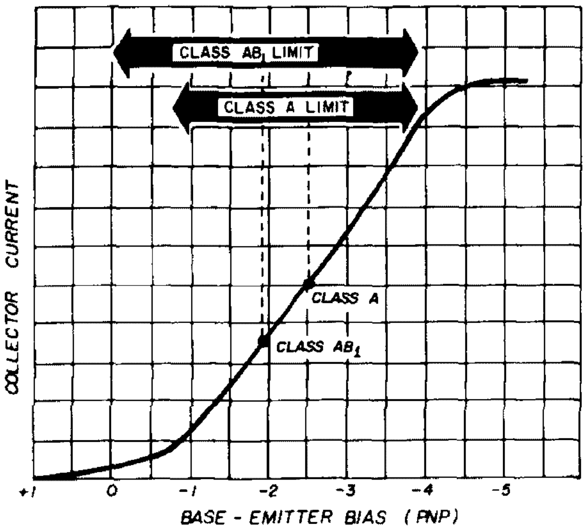


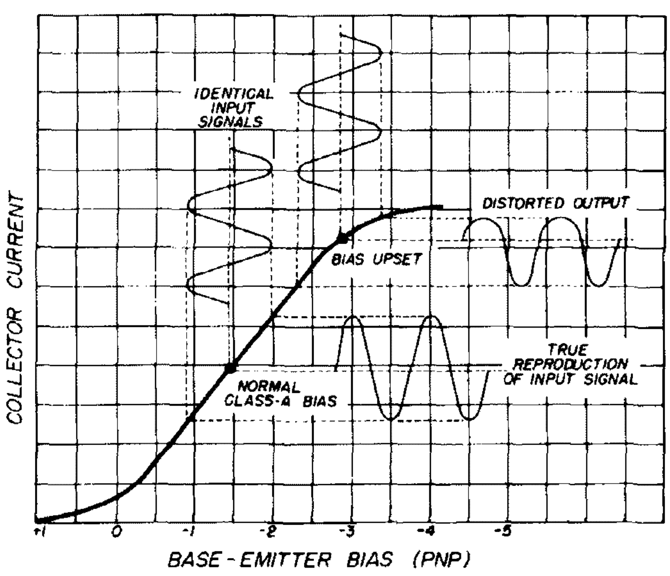

**fig. 1. Base-collector transfer curve of a typical pnp transistor. Class-A operation is entirely linear. Class AB1 has bias nearer cutoff, but signal input is limited so it never actually reaches the cutoff or saturation portions of the curve.**

tened at the knee of an amplifier curve.

## the square test

Your best instruments for distortion-chasing are the square-wave generator and oscilloscope. The units used for this article are illustrated in **fig. 3**. The generator is a kit-type and inexpensive—plenty okay for ham testing. (You can use it, and the techniques I'm about to describe,

**fig. 2. A class-A amp distorts input waveforms if the bias shifts. Here, severe compression of the positive half-cycles (of the output) is caused by the curve's knee.**




for hi-fi testing, too.)

Why not a use sine wave, you wonder? The reason is simple. It's hard to spot slight distortion in a sine wave. In a bad case, yes, the twisting or flattening is visible. But you just can't see subtle alterations except by superimposing input-output waveforms with a dual-beam scope or an electronic switcher. And that's more expense and complication.

So use a square wave. Feed a square wave into your receiver and transmitter voice stages and look at the waveforms you get with your own scope. Become familiar with them.

If you have a Polaroid camera, take pictures of the scope patterns when the equipment is operating normally. (Do it in the dark, and use ASA-3000 Polaroid film.) On the back of each picture, mark the brand and model of the equipment, what test point the scope is connected to, and the normal amplitude of the waveform. Keep the photos with the service manual and schematic.

There are two nominal square-wave frequencies for testing a communications audio amp. Bandpass of the stages is limited, remember. Low frequencies seldom go below 100 Hz, and highs seldom much above 3000.

A square wave stays square on the *trailing* edge if low frequencies are okay down to one-tenth of the fundamental frequency. Therefore, a 100-Hz waveform should stay square at the trailing (right-hand) corners if the stage it's going through has normal response down to 100 Hz. The normal input waveform in **fig. 4A** takes on a tilt like **fig. 4B** if the bass response is limited more than it should be.

On the other hand, the *leading* edge of the square wave tells about the high frequencies. The leading corner of the waveform stays square if frequencies are okay up to ten times the fundamental frequency. With a 3000-Hz upper limit in communications gear, a 1000-Hz square wave would show rounding at the leading edge—even in a normal amplifier. You can see this rounding in **fig. 4C**.

What you do is move the generator frequency to 300 Hz. The *leading* edge of the square wave is the one to watch. Just ignore the trailing edge, because it now tests down to 30 Hz—well below the amplifier's normal response. It'll be rounded. The waveform in fig. **4D** shows the amplifier being tested is good to 3000 Hz. The leading edge is square.

## distortion on the square

A few unusual distortion waveforms are pictured in fig. **5**. Two of them illustrate faults that cause distortion but are not necessarily from a nonlinear amplifier. In their cases, the amplifiers were self-oscillating.

In fig. **5A** you can see a parasitic oscillation modulated on the square wave. Also obvious, once you know to watch for it, is the rounding on the leading edges—the sign of poor high-frequency response. This speech-amp trouble was traced to a bad bypass capacitor. It was letting the stage oscillate and also overload on below-normal signal levels. You should have heard the crazy sound this transmitter produced!

The waveform in fig. **5B** reflects an extreme case of self-oscillation. The square wave was triggering the oscillation, one burst for each cycle of square wave. Voice signals were unintelligible. Yet, a sine wave showed hardly any distortion. It took a waveshape with a sharp "attack" to set the stage oscillating. Voice signals were plenty sharp enough.

Bias distortion in an amplifier stage created the terrible waveform in fig. **5C**. This one was a fooler. Tracing a square wave through the stages with a scope uncovered fairly quickly the driver that was distorting the signal. But the actual defect turned out to be a bad transistor in an earlier stage. Coupling was dc. The leaky transistor fouled up the bias on the driver, yet the scope showed the square wave going through the defective transistor okay. It just happened that bias for the driver was more critical than in the earlier stage.

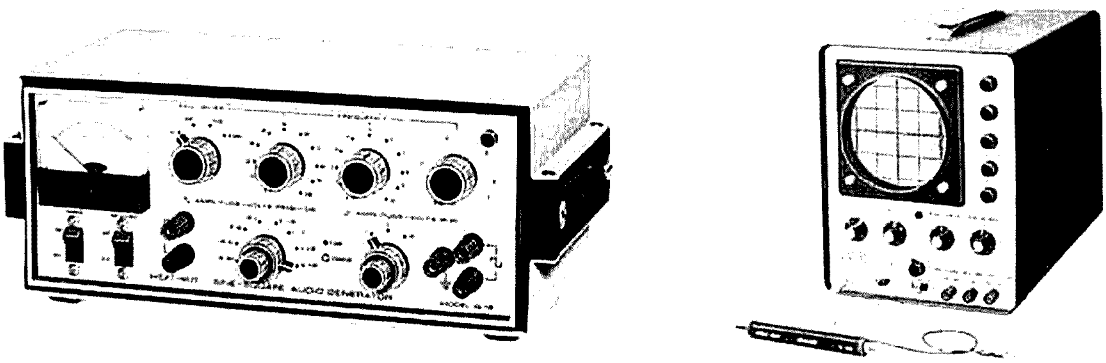


**fig. 3. Scope and sine-square generator used to test the amplifiers for this article. Both are service-type instruments—not lab equipment.**

A bad capacitor created the severe overload waveform in fig. **5D**. You could duplicate this waveform by simply feeding too much signal through any stage. But in this one the input signal was well below the natural overload level. The stage just couldn't handle much signal. A leaky coupling capacitor had drastically altered bias on the transistor.

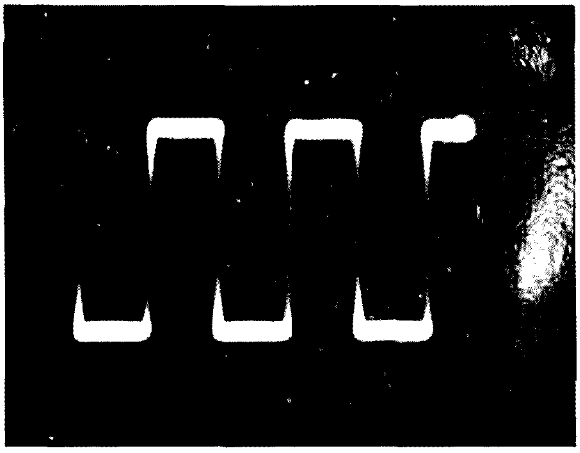
You'll see plenty of other distortions as you test squawky-sounding amplifiers with square waves. You don't always have to recognize the trouble from the waveform, although practice will make you pretty good at doing that. What the generator and scope do is identify the stage where the distortion is. Then, you track the defect down by specific troubleshooting within the faulty stage.

Of course, the end goal is to find the defective part. Once you've identified the stage that's bad, there aren't usually many parts to test. But it's also generally quicker to be scientific about it, rather than haphazard.

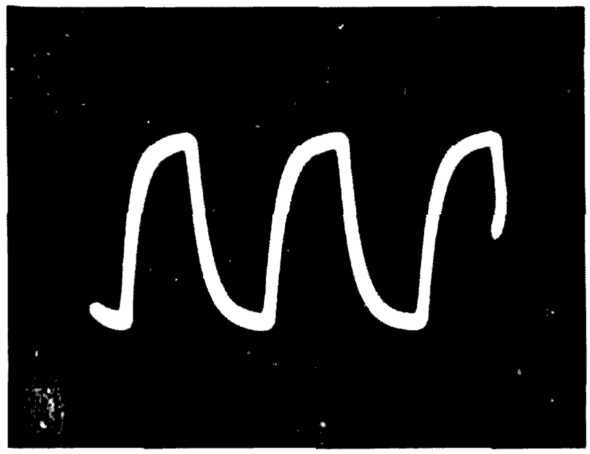
Using the typical speech amp *in fig. 6* as an example, I'll tell you the steps I go through to find the bad part. It works no matter what the type of distortion, so just what the waveform looks like isn't that significant.

First disconnect the mike. Connect the square-wave generator in its place. Turn the generator up high enough to simulate the mike input. The average crystal or ceramic mike puts a millivolt or two into a high-impedance input.

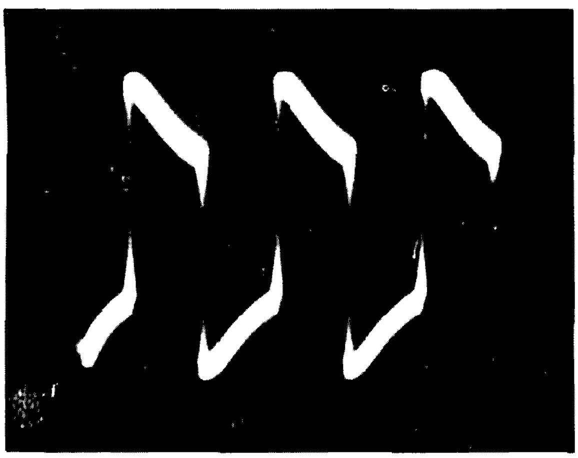
If you want to be sure, connect the scope across the microphone input, with the mike in place. Say "ahhhh" into the mike at close range and notice the amplitude of scope trace. Then unplug the mike and substitute the generator. Turn up its output control until the scope display is the same height it was with the mike. The scope should show you a waveform like fig. 4A.



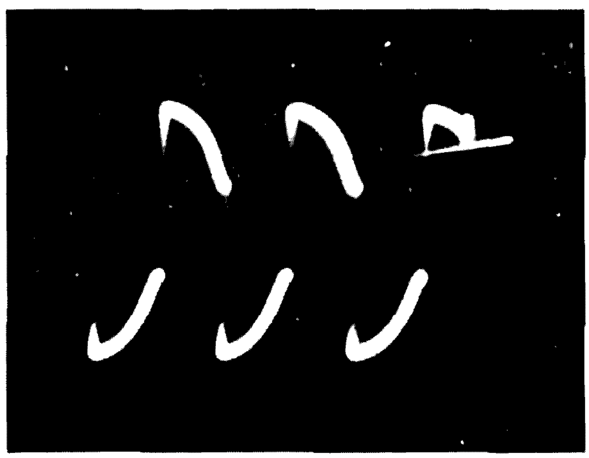
**A. Normal, generator directly to scope**



**C. Rounded leading edge, poor high response**



**B. Tilted trailing edge, poor low response**



**D. Normal, when 300-Hz square wave is used; amplifier doesn't go down to 30 Hz, hence downturned trailing edge**

**fig. 4. Square waves are informative for testing.**

Then move the scope from test point to test point. Keep in mind what I already told you about the frequency response of the speech amps. The further along you get, the more rolloff you find above 3000 Hz. It's especially pronounced after Q2 (C8 bypasses a lot of highs). Check both ends of the response; use 1000 Hz for the low end and 300 Hz for the high end. If that sounds strange, reread the paragraphs that describe "the square test."

Make the tests in sequence. A bad waveform at A tells you R1 or C1 is in trouble. Trouble at B is most likely the fault of C2 or C3—or perhaps R2. However, a defect in Q1 could reflect back. For example, a base-emitter short could load down the waveform at point B.

A bad waveform at C points to Q1 or C4. At D it could mean C5 or R5 is bad. Don't forget to try turning the shaft of R5, though, to make sure it's not just turned down.

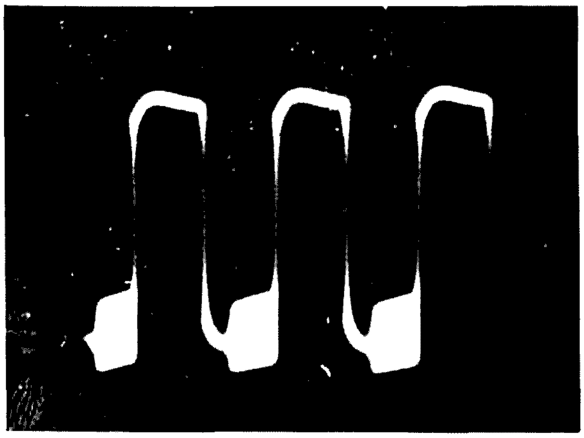
And so on. Each test point tells you about the components connected there and those between there and the test point just preceding. You narrow down the fault to just a few parts, which you can then test individually without much wasted time.

Pay attention to the kinds of stages. You might be misled if you don't. Q1 and Q3 in **fig. 6** are common-collector amplifiers (or emitter followers, if you prefer). They don't impart any voltage gain to the square-wave signal. However, Q2 does, and you should see a considerable increase in amplitude between points E and F.

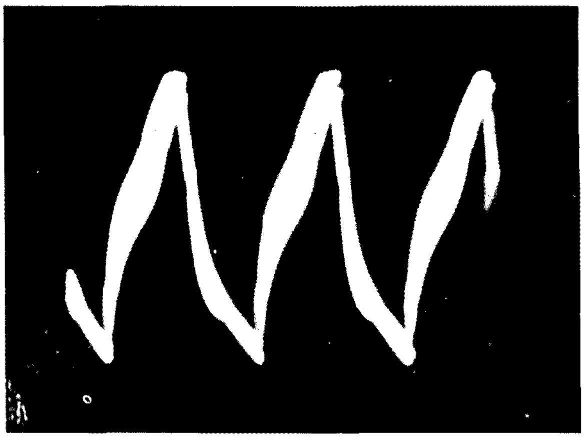
## botched-up bias

When you find trouble around a transistor, don't just jump in and replace it. Measure the dc voltages first. Distortion, as you read earlier, is commonly the fault of wrong bias on a transistor. True, it may be due to transistor leakage, but it could just as likely be a bad resistor or a leaky capacitor.

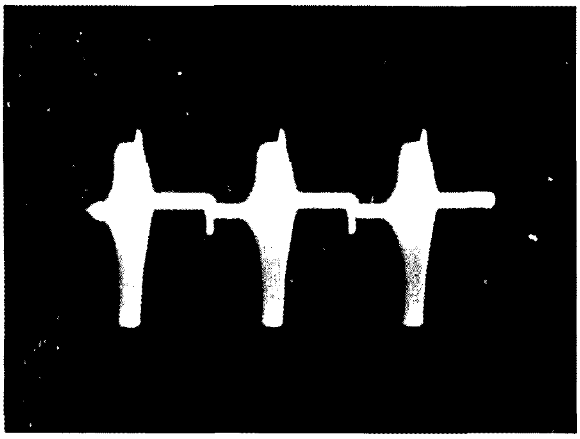
For example, suppose in a system like that diagramed in **fig. 6** you touch your scope probe to point D and find the square wave still normal. At point E there's a tiny change but not enough to worry you yet. But the waveform at F is lousy. The trouble seems isolated between E and F. Is Q2 defective? That's



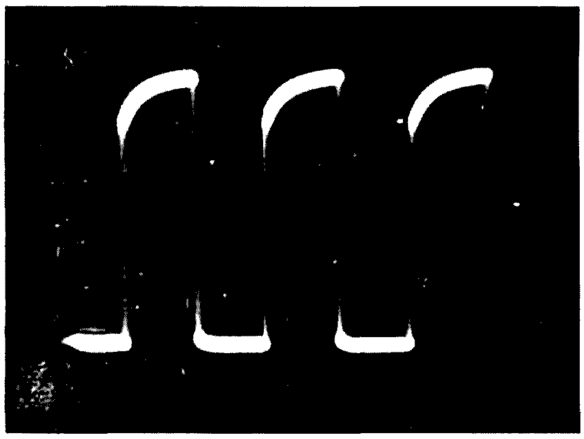
**A. Parasitic oscillation**



**C. Distortion makes sawtooth**



**B. Oscillation, amplifier blocked**



**O. Overload distortion**

**fig. 5. Waveforms that reveal distortion.**

usually your first thought.

If you have a transistor tester, great. Use it, and save yourself a lot of measuring and figuring. But you probably haven't, so you go ahead with voltage measurement.

Collector voltage is wrong, but so is base voltage. Strangely, the base voltage changes as you turn modulation level control R5. That alone should tell you what's wrong. But maybe you don't happen to try turning R5. You check the two most likely parts: R6 and R7. Both measure okay with your ohmmeter. You disconnect the transistor, just in case the base is shorted to *emitter or collector;* normally there is hardly any base current to drag the base voltage down.

When you disconnect C6 to test it, the base voltage goes back to normal. The capacitor is very leaky. You can test it with your ohmmeter, or by the "open end" test (disconnect the end nearest ground and measure for voltage leaking through from the base-bias circuit). The leakage didn't affect the waveform between points D and E, but it sure did play heck with the bias on Q2.

Watch out for bad bypass capacitors. They're hard to spot, especially if they're open. Capacitors like C1, C3, C4, and C8 show definite trouble when they short, or even if they get very leaky. The fault may be a little harder to spot if one of them is open.

The best way to check them is with the scope connected at the output (toward the balanced modulator). Just bridge each capacitor with a good one of the same value while you watch the square wave. If the waveshape changes, the one you bridged is open. Watch the corners, particularly at the leading edges. These bypasses affect *high-frequency* response.

An emitter bypass like C7 has more effect on low frequencies. With the scope connected as just mentioned, bridge the capacitor with a known good one. If the trailing corner of the square wave alters much, replace the capacitor. It's probably open. If it were leaky, it would change the emitter voltage drastically.

A sensible combination of bias checking and square-wave tracing should solve almost any distortion trouble for you. At least, it gets you so close to the root of the trouble it leaves only a few parts to test.

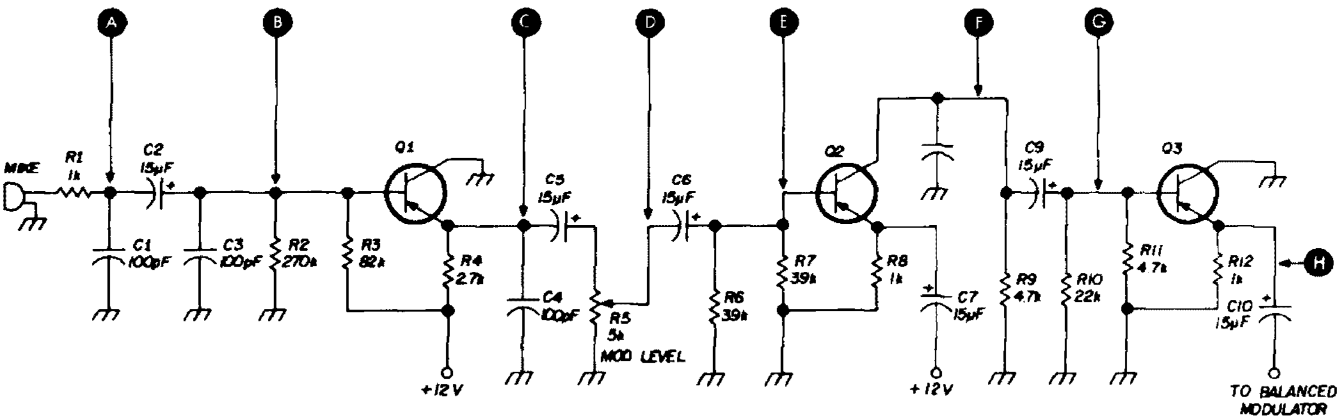



**fig. 6. Points for square-wave testing in a typical speech amplifier. Voltage or substitution tests find the defective part.**

## on the griddle

For my next column I've cooked up some solutions to a problem that aggravates even experienced professional servicers. Troubleshooting rf and i-f stages has peculiarities all its own. Yet, you'll be surprised how simply it can be done. All you need is a little advance knowledge of what to expect, and some hints for what tests to make. Both are in **repair bench** next time.

**ham radio**

# an improved six-meter converter

A new approach to portable vhf converters using fets and a tunable local oscillator

Rick Littlefield, K1BQT, 209 Marin Street, San Rafael, California 94901

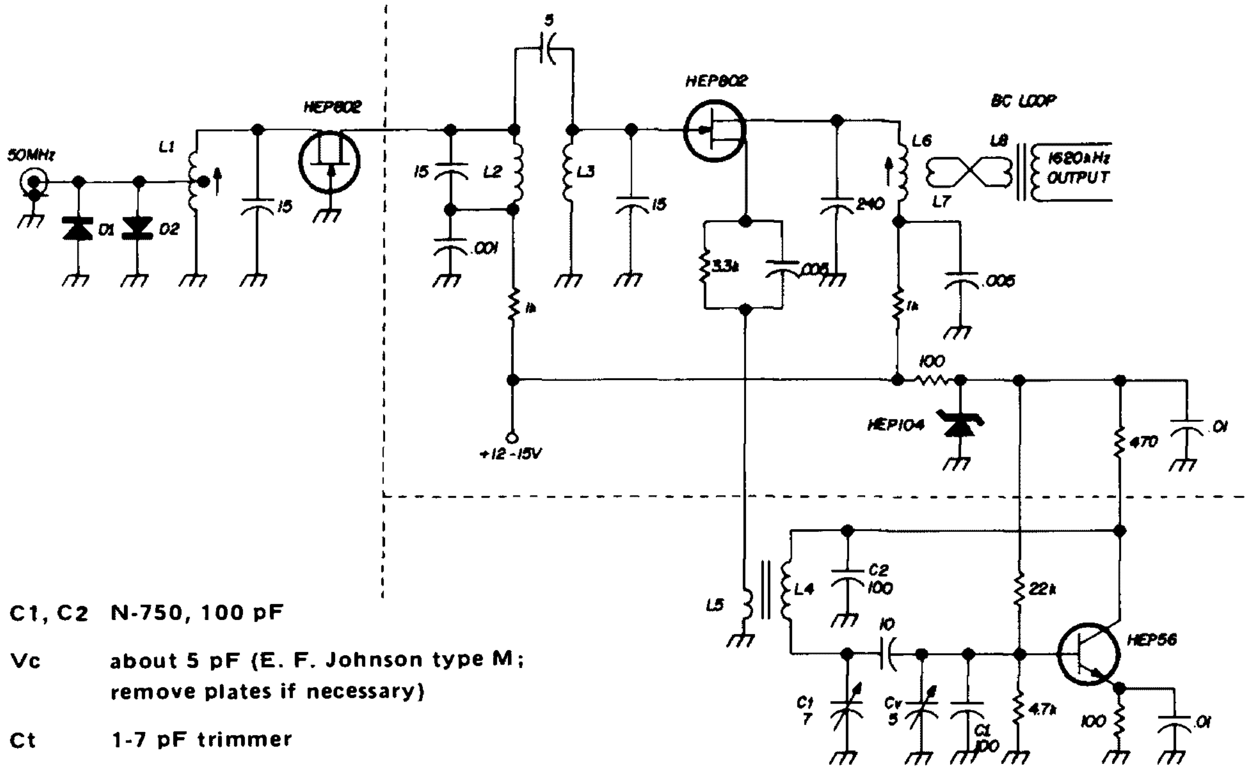
**Many converter circuits** have appeared in print for those who wish to use an inexpensive transistor broadcast set in a solid-state receiver or transceiver. I've built a few of these converters, and in many cases they turned out to be a compromise between simplicity and performance. So I designed one that's simple but free of a few traditional flaws. The resulting circuit uses inexpensive fet's in rf and mixer stages and a Vackar tunable oscillator for frequency control (**fig. 1**). This approach represents an improvement over the new autodyne and crystal-controlled bipolar circuits.

## design

The tunable-oscillator-type converter offers the best design approach, because the transistor radio can be used to maximum advantage as a fixed-frequency i-f amplifier. By setting the bc receiver to a frequency just above the broadcast band, 1620 kHz, image rejection is held to the highest possible level over the six-meter band, and local broadcast-station feedthrough is eliminated. Also the mixer output can be peaked with a high-Q circuit for higher gain. None of this is possible when the transistor radio has a tunable i-f.

The most obvious drawback to this approach is the requirement for a stable, tunable oscillator in the 50-MHz region. To solve this problem I chose a solid-state Vackar oscillator circuit. The version in

this converter uses an easily obtainable toroid inductor,* and provides more stability than any simple circuit I know of. If good vfo construction practice and the prescribed temperature compensation are weakest signals. The circuit is very simple to build and adjust, since it tunes quite broadly and is not prone to oscillation. If more gain is required, the optional amplifier shown in **fig. 2** provides a hotter




| | |
|---|---|
| **C1, C2** | **N-750, 100 pF** |
| **Vc** | **about 5 pF (E. F. Johnson type M; remove plates if necessary)** |
| **Ct** | **1-7 pF trimmer** |
| **L1, L2, L3** | **0.55-0.80 mH (Millen 69054-0.68)** |
| **L4** | **9 turns no. 22 on T50-10 core (see text)** |
| **L5** | **2 or 3 turns test probe wire (depending on oscillator output) on cold end of L4** |
| **L6** | **20-50 µH (Calectro D1-854)** |
| **L7** | **20 turns no. 30 scramble-wound slightly above L6 on same form** |
| **L8** | **5 or 6 turns no. 22 on cold end of bc loop antenna, connected to L7 with twisted hook-up wire** |
| **L9** | **10 turns no. 24 tapped 3 turns from gnd on ¼" slug-tuned form (Miller 4500-4)** |

**fig. 1. Schematic of the fet six-meter converter.**

used, you'll have no trouble with instability.

The rf and mixer have a low noise figure and a much wider dynamic range than is possible with bipolars. Similar circuits can be found in the more recent editions of the ARRL handbook. The grounded-gate rf stage, while not an extremely high-gain circuit, does provide sufficient gain for reception of all but the front end. More care must be used in layout and tuning procedure, because the alternate circuit tunes quite sharply and requires neutralization. The ARRL handbook provides tuning procedures for grounded-source amplifiers, should you run into a problem.

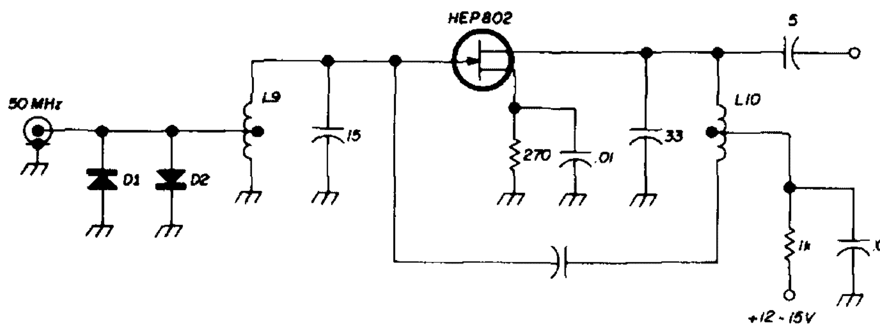
The double-tuned circuit between the rf and mixer is a worthwhile investment. It improves out-of-band signal rejection. The mixer is a standard grounded-source circuit, with oscillator injection via a grounded link. The output is tuned slight-

*A kit of two toroid forms and wire is available from Circuit Specialists Co., P. O. Box 3047, Scottsdale, Arizona 85251 for $1.25. Order: VFO-6.

ly above the edge of the broadcast band and is fed to the transistor receiver loop antenna by a simple balanced line.

## construction

Component arrangement is somewhat flexible, depending on the size of your circuit board and your needs. My board measures a scant 1¾ x 3¼ inches; even this is too small unless you enjoy building ships in bottles.




**L9 0.55—0.80 mH (Millen 69054-0.68)**

**fig. 2. Alternate 50-MHz rf amplifier provides more sensitivity than the grounded-gate stage used in fig. 1 but must be neutralized to avoid oscillation.**

A few considerations should be kept in mind when laying out and building the oscillator. I mentioned earlier that some form of temperature compensation is needed. This can be provided by using N750 capacitors for C1 and C2. Also, to prevent overcoupling between oscillator and mixer, the grounded link is made from rubber-insulated clip-lead wire. The insulation thickness allows fairly loose coupling, yet provides a snug support for the link. The coil assembly is then secured with epoxy cement.

Finally, if you operate the transistor radio and converter from the same battery, be sure to include the zener regulator. Otherwise, the varying load of the bc radio's class B output stage will cause the oscillator to drift.

## adjustment and operation

By listening on any good six-meter receiver, you can find the oscillator signal. Then set the signal for the tuning range you wish. Mine tracks from 51.5 to 52.7 MHz, which covers only the populated low end of the band. A noise generator should be used to tune the mixer, since the mixer pulls the oscillator frequency just enough to make signal-generator peaking a frustrating experience. I used a regenerative receiver, placed across the room as a noise source.

Once properly tuned, the unit's sensitivity should produce a noticeable increase in background noise when an antenna is connected. The signal-to-noise ratio may be optimized by tuning in a weak signal and adjusting the rf-input coil for the desired characteristic. With the grounded-gate amplifier, little difference between maximum gain and maximum s/n will be noticed; however, with the grounded-source circuit, the significant improvement can be achieved.

## performance

The performance of my unit is more reminiscent of a modest communications receiver than of the shoestring converter/bc set configuration. Sensitivity is excellent, and you may want to broaden the i-f response of the transistor radio if its selectivity seems too sharp for your operating needs.

Although overall gain won't blow you out of the room, the low noise figure allows you to hear weak signals that would be buried in noise of a similar bipolar setup. The circuit is well worth considering if you're planning a portable rig or are tired of fighting the drift problems and low sensitivity of older designs.

**ham radio**

# improving the intelligibility of communications receivers

Bill Rogers, WA5RAQ, 1121 Southwest 53rd Street, Oklahoma city, Oklahoma 73109

**The sound reproduction clarity** of communications receivers can be improved significantly by simply improving the speaker enclosure. Common utility speakers are poor performers at best and can be improved (or impaired) by the baffles they are mounted in. Much of the acoustic distortion is caused by either speaker resonance, cabinet resonance, cabinet reflections, back waves or poor frequency response, and these can be easily corrected with a good enclosure. We have all become accustomed to the sound of a small speaker in a metal cabinet or the boomy sound of a large speaker in a box just big enough for it, but these systems leave a lot to be desired.

Sound reproduction can be improved by using a small high-fidelity speaker or hi-fi headphones. The small hi-fi speakers I use on ssb and cw have resulted in a big improvement in communications effectiveness.

## how to do it

Use a small round or oval speaker, 5-inch or 5x3-inch, mounted in a wooden box made from ¾-inch pine or plywood. The box can be just big enough to contain the speaker with 2 or 3 inches of depth. Use 1/8- or 1/4-inch Masonite or plywood for the front panel. Cut the speaker opening out completely—don't drill a few small holes in a cute pattern and hope for the best. Cover the front panel with screen, an open grill with very small reflecting surfaces or plastic grill cloth designed for hi-fi speaker systems.

Mount the speakers and grill so there is no vibration of loose parts. Then fasten thick sound-absorbent material such as felt or foam rubber on the inside surfaces of the box. When the speaker enclosure is completed, put it near the operating position, preferably with the center axis of the cone directed toward your head. If your receiver has an internal speaker it should be disconnected.

## how it works

The thick wood of the enclosure reduces cabinet vibrations, and the unrestricted speaker opening practically eliminates reflections between the cone and grill. The absorbent material helps to damp speaker response and reduces reflections between the inside walls, resulting in smoother frequency response. The small speaker and small box reduce low-frequency response; boominess is reduced by the cancellation effect of the back-wave.

With the speaker placed close to the operating position the volume control can be turned down. This reduces undesirable sound reflections from nearby objects and reduces distortion caused by any non-linearity in the speaker mechanism.

**ham radio**

Reprinted with permission from the *"Collector and Emitter,"* club publication of the Aeronautical Center Amateur Radio Club, Oklahoma City.

# quad-antenna design parameters

W. R. Hillard, K6OPZ, 28700 King Arthur Court, Palos Verdes Peninsula, California 90274

**The extensive use of** parasitic antennas over the past thirty years or so has resulted in much published data. Design information is available that allows the amateur to easily construct two- or three-element monoband parasitic antennas with little or no experimenting or tuning, except perhaps for feed-point matching.

The quad antenna, which is a special type of parasitic array, unfortunately has been neglected as far as detailed design information is concerned. However, when it is recognized that the quad exhibits the characteristics of conventional parasitic arrays except for differences in element length caused by the folded wire elements, quad design is very similar to that of conventional parasitic arrays and is not particularly mysterious.

This article presents a design approach that allows construction of a quad antenna with no special element tuning or fussing with element lengths, tuning stubs, or loading coils.

## characteristics

Before proceeding with details of the quad design approach, it is well to discuss some facts regarding the quad and dispel some misunderstandings.

**forward gain.** Quad antenna gain has been underrated in some of the literature. If a quad loop is considered as one element, its gain can be shown to match or exceed that of a monoband Yagi on an element-for-element basis.

**front-to-back ratio.** The quad really shines here. Its front-to-back ratio will exceed that of a Yagi with the same design criteria by up to 10 dB (based on element spacing, element length, and related factors).

**vertical radiation angle.** The vertical radiation angle for any horizontally polarized antenna is a function of its height. Therefore, any half-wave dipole, Yagi, or quad will have the same vertical radiation angle if they are all at the same height above ground. The quad's performance as an excellent band opener and closer is due to its directivity characteristics, not to its vertical radiation angle.

**element interlacing.** The minimal mechanical and electrical complications of accommodating three bands on one antenna structure with no apparent loss in efficiency is a definite advantage of the quad.

**bandwidth.** The bandwidth of an antenna can be based on (a) the frequency range beyond which the feedpoint vswr exceeds a specified value, usually 1.7 to 1; (b) the bandwidth over which front-to-back ratio or gain does not drop below a certain value; or (c) the point at which either the director or reflector closely approaches resonance. Because the director and reflector are electrically close to the driven-element length (much more so than with a tubing Yagi), the point at which the director or reflector approaches resonance will govern the bandwidth of a quad designed for optimum performance.

**rain static.** Compared with the Yagi, the quad is relatively immune to rain or snow static.

**cost.** Prices of aluminum tubing in the United States are low compared with most countries. In addition, a large variance exists in prices of aluminum bought from surplus sources versus those of new aluminum bought from metal suppliers. The price of a home-constructed Yagi therefore depends on how good a shop-

ping job is done. It is usually found that a well-built, home-constructed quad with fiberglass cross arms will cost more than a monoband Yagi. This is also true of purchased quads. However, quads will tolerate a lighter rotator and tower than a comparable Yagi because of the quad's lighter weight. Therefore, the quad is usually a less expensive total antenna system.

A word of caution: aluminum cross arms are not satisfactory for a quad, since they degrade antenna performance and absolutely ruin front-to-back ratio. Breaking up the aluminum cross arms with insulators helps, but this is difficult to implement.

with two or three loops of safety wire to anchor the wire to the quad arm, as shown in **fig. 1.*** Start with the element too long, with a slight droop and a temporary splice at the bottom. Note that quad parasitic elements not using stubs or loading coils are a continuous wire loop. The element length can be adjusted for resonance, maximum gain, or front-to-back ratio. Once the proper length is obtained, the loop splice can be permanently soldered, coated with silicone grease, and the loop run outboard (but equally spaced) to tighten the loop into a perfect square. Another method is to drill a hole in the arm several inches outboard from where the element would

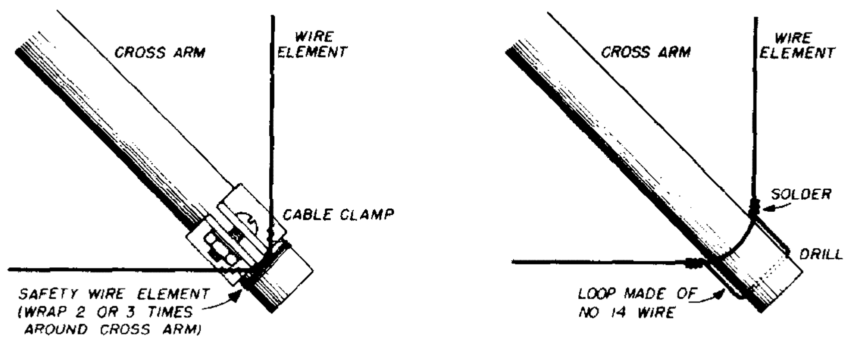



**fig. 1. Two methods of securing quad elements to cross arms.**

## stubs and loading coils

Since the quad is a closed loop, adjustment of element lengths is difficult. Adjustment of element lengths is usually done with tuning stubs or loading coils. Stubs are unsightly, and the problem always arises of how to anchor them so they don't move in a wind. Both stubs and loading coils destroy the electrical symmetry of a quad loop. All in all, stubs and loading coils are best avoided.

Alternatives are to (a) use element lengths established by criteria such as those in the following paragraphs and not attempt to tune the quad elements, or (b) select the optimum element-loop length on a trial basis.

## element length adjustment

There are two ways to adjust element-loop length. Instead of drilling holes in the bamboo or fiberglass arms, small nylon or metal cable clamps can be used, normally be anchored. Then use a wire loop to hold the quad loop in place. Slack in the quad loop created by element adjustment can be accommodated by the small loop.

## boom length

Factors that govern boom length for a Yagi apply also to a quad. Most quads are built with a boom too short to realize maximum performance on 20 meters, which is a highly competitive band. To achieve a good balance between practicality and performance, a 20-meter quad should have a boom length as follows:

| | |
|---|---|
| Two-elements | 10 feet |
| Three-elements | 18 to 20 feet |
| Four-elements | 24 feet |

As with a Yagi, placing more than one

*The safety wire should be copper to avoid galvanic action, which results in corrosion. **editor**

table 1. Element dimensions for a quad with a 14-foot boom.

| | 20 meters | 15 meters | 10 meters |
|---|---|---|---|
| reflector | 2.8% long (1.028)(70.7) = 72.68 ft | 2.1% long (1.021)(47.25) = 48.24 ft | 2.1% long (1.021)(35.1) = 35.84 ft |
| driven element | $\frac{1004}{14.2}$ = 70.7 ft | $\frac{1004}{21.25}$ ≈ 47.25 ft | $\frac{1004}{28.6}$ = 35.10 ft |
| director | 1.4% short (.986)(70.7) = 69.71 ft | 1.2% short (.988)(47.25) = 46.68 ft | 2.1% short (.979)(35.1) = 34.36 ft |

director on a 20-foot boom just doesn't give a satisfactory return for the effort *on 20 meters.* Definite advantages can be obtained with quads that have more elements on 10 and 15 than on 20. For example, an effective quad is one with a 20-foot boom having three elements on 20 and four on 10 and 15 meters.

## boomless quads

Three-band, two-element quads using no boom and the same element spacing on 20, 15 and 10 meters have only one real advantage over conventional quads: all can have the same feed-point impedance. A conventional quad has its widest element spacing on 10 meters, which helps to broadband the antenna on this band.

## design graph

I developed the graph shown in **fig. 2** expressly for high-frequency quad arrays. Using the graph and the approach shown in the following design example, a quad can be built with no stubs, loading coils, or tuning. For those who wish to avoid the math involved in the design example, **table 1** was developed from **fig. 2** for popular quads that can be purchased or homebuilt.

Design graphs similar to that in **fig. 2** have been published for Yagis. Because of its wire elements and the fact that the loop is, in effect, two half-wave dipoles folded at the ends and stacked a quarter wavelength, the quad has characteristics that make its element dimensions much different from those of a tubing Yagi array. These factors, plus others, were considered in the analysis used to develop **fig. 2**, which expressly applies to a quad designed for 20, 15, and 10 meters.

## design example

As an example of using **fig. 2**, suppose it is desired to design a three-element quad for 20, 15, and 10 meters on a 14-foot boom with 6-foot director spacing and 8-foot reflector spacing. The desired center frequencies are 14.2, 21.25, and 28.6 MHz. Proceed as follows:

**1.** Establish percentages for the desired band excursion within which the antenna must be effective. To cover 20 meters, the excursion from 14.2 to 14.0 MHz is 200/14.2 = 1.41 percent, and the excursion from 14.2 to 14.35 MHz is 150/14.2 = 1.056 percent. For 15 meters, the excursion is 1.175 percent to reach 21.45 MHz, with a 21.25-MHz center frequency.

Ten meters represents a problem requiring a compromise. To cover the entire band, the excursion is 2.1 percent to reach 28.0 MHz and 3.84 percent to reach 29.7 MHz. The parasitic elements must be so far from optimum to avoid resonance that performance is severely compromised. A good alternative is to settle for ±2.1 percent, or 28.0 to 29.12 MHz. The phone man may elect to choose 29.1 MHz as center frequency for 10 meters.

**2.** Establish element spacings in wavelengths for each band using the formula:

Element spacing is $\lambda = xf/984$, where x is the element spacing in feet and f is the design-center frequency in

MHz. Using the formula, element spacings are:

| band | director (6 ft) | reflector (8 ft) |
|---|---|---|
| 20 meters | 0.086 | 0.115 |
| 15 meters | 0.13 | 0.172 |
| 10 meters | 0.175 | 0.234 |

3. Identify element spacing (fig. 2).

4. Select parasitic-element length as a percentage of driven-element length, bearing in mind the reactance excursion of the element across the band. Note that a reflector spaced 0.1 wavelength from the driven element should be 2.8 percent longer than the driven element. A director spaced 0.15 wavelength from the driven element should be 1.4 percent shorter than the driven element. Shown in fig. 2 are the choices I made using this technique (circled values).

It's not always possible to choose optimum element lengths at the design center frequency and avoid parasitic-element resonance at the band edges. In this regard, it's sometimes better to provide wider spacing for the director than for the reflector.

5. Calculate element dimensions. The driven element length equals 1004/f in MHz. This is 70.7 feet for 20, 47.25 feet for 15, and 35.1 feet for 10 meters. The reflector for 20 meters should be 2.8 percent longer than the driven element, or (1.028) (70.7) = 72.6 feet. The reflector for 15 meters should be 2.1 percent longer than the driven element, or (1.021) (47.25) = 48.24 feet. Calculation of each element results in the data listed in **table 1**.

When the element lengths are established using the above approach or by direct use of **table 1**, all that remains is to construct the array and decide on the feed method. The data in **table 1** are applicable to a two-element quad with 8-foot reflector spacing by simply ignoring the director data. The data are applicable to a four-element quad on a 20-foot boom. Make the first and second directors the same length.

## tuning

The above design approach should not be considered an absolute cure-all that eliminates benefits from tuning the array.

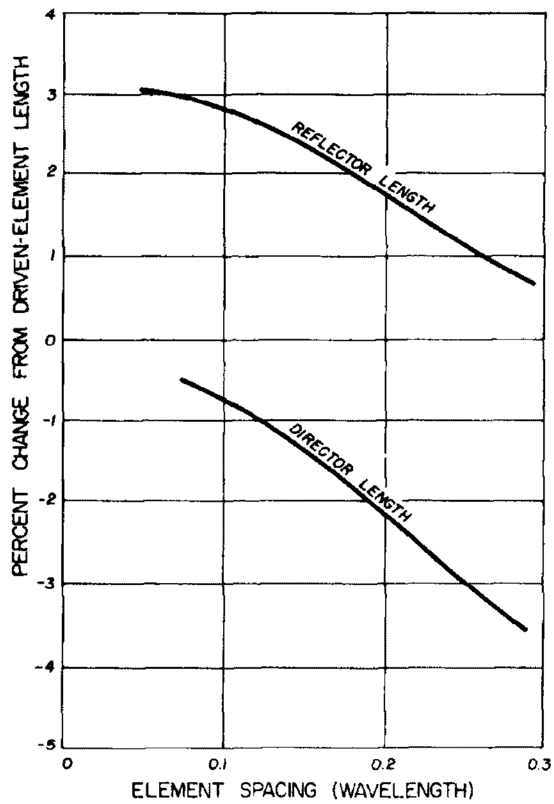



**fig. 2. Optimum director and reflector lengths as a percentage of driven-element length as a function of element spacing.**

However, few have the talent, equipment, facilities, or perhaps patience to properly tune an array, particularly one with three or more elements. For these amateurs, the "no tuning approach" will be most useful, and those who wish to tune by experimenting with loop length will find that this article defines an excellent starting point. If you wish to tune the quad, using the preceding design data as a starting point, some comments are in order.

The curves in **fig. 3** are recommended element dimensions. At design center frequency, maximum gain will usually be found with a dimension closer to the driven element length than that shown in **fig. 3**. Correspondingly, maximum front-to-back ratio will usually be found with a dimension shorter for a director and longer for the reflector than shown in **fig. 3**. The design must be compromised to achieve best overall performance consistent with gain, front-to-back ratio, and bandwidth.

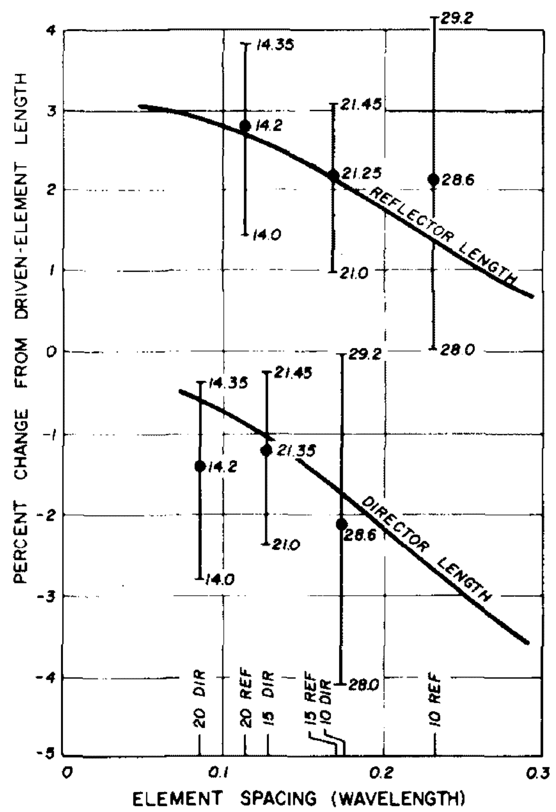



**fig. 3. Plot showing the use of fig. 2 in establishing quad element lengths for the design example discussed in the text.**

Tuning for maximum gain at the design center frequency will usually cause narrowband response. If the array is tuned for maximum forward gain, check the response at the band edges to get the best overall gain across the band. As an example of how dangerous forward-gain tuning can be, a director spaced 0.1 wavelength from the driven element will theoretically give maximum gain when it is the *same length* as the driven element, and a reflector spaced 0.25 wavelength from the driven element will theoretically give maximum gain when it is the *same length* as the driven element! Such conditions are obviously bad, because a small frequency excursion can convert the director to a reflector, or vice versa. If you tune for forward gain, try for good response across the whole band, or at least over the band segment of interest.

**table 2. Typical quad antenna feed methods.**

| antenna | quarter-wave matching transformer | main feedline |
|---|---|---|
| **2-element quad** | | |
| 10 | none (direct feed) | RG-63/U |
| 10 | two pieces RG-63/U in parallel (electrical length ¼ λ ea) | RG-8/U<br>RG-58/U |
| 15 | none (direct feed) | RG-59/U<br>RG-11/U<br>RG-62/U |
| 15 | one piece RG-11/U or RG-59/U (electrical length ¼ λ) | RG-8/U<br>RG-58/U |
| 20 | none (direct feed) | RG-8/U<br>RG-58/U |
| **3-element quad** | | |
| 10 | none (direct feed) | RG-11/U<br>RG-59/U |
| 15 | none (direct feed) | RG-11/U<br>RG-59/U<br>RG-8/U<br>RG-58/U |
| 20 | none (direct feed) | RG-8/U<br>RG-58/U |
| **4-element quad** | | |
| 10 | | |
| 15 | none (direct feed) | RG-8/U |
| 20 | | RG-58/U |

Tuning for front-to-back ratio is safer and usually much more effective. Such tuning is, however, very touchy. Because of nearby objects it can be frustrating and sometimes leads to inconsistent results.

Don't attempt to tune an array unless the services of a nearby amateur are available, or you can locate a temporary transmitter site several wavelengths and preferably at least a mile from the array. Use the nearby amateur or your remote site for transmitting, and use a receiver with an s-meter or field-strength meter at the site of the array being tuned.

## feed-point matching

Several techniques for feeding the quad can be used. I prefer the gamma match with separate feed lines, but any of the following methods are effective.

The simplest method is to connect all three driven elements together, using a single insulator across the 15-meter driven element. The system can be fed with a single 52-ohm line. This is an effective feed method despite published statements to the contrary. Measurements with an impedance bridge have shown a good match on all three bands.

A second approach is to use one of the methods in **table 2**. These are also effective, as proved by countless users. Some mismatch will occur, however.

## the gamma match

The preferred matching system for quad antennas is the gamma match. It's simple, easy to use, and reliable. Referring to fig. 4, the gamma match is implemented and tuned as follows:

1. The coax cable shield connects to the exact center of the driven element. No center insulator is used.

2. Capacitor C is housed in a plastic box. A drain hole should be drilled in the bottom. The box should be sealed with cement. It can be secured to the driven element with a clamp.

3. The matching stub is made of the same wire as the quad element, usually no. 14 copper wire. Spacers can be pieces of polystyrene or any good plastic rod.

4. The values of capacitor C and dimension A are as follows:

| band | capacitor C | dimension A |
|---|---|---|
| | (pF) | (inches) |
| 10 | 50 | 12 ±6 |
| 15 | 100 | 15 ±8 |
| 20 | 140 | 25 ±10 |

5. Tune as follows: connect a Monimatch or similar instrument, driven by your exciter, to the antenna feed point (fig. 4B). Adjust dimension A and vary capacitor C for minimum standing-wave ratio.

## conclusion

In the final analysis, antenna selection should be based on many factors, including the antenna's adaptability to your location and operating habits. There is no "miracle antenna." The size of a well-designed antenna is a good guage of its expected performance. In general, if you

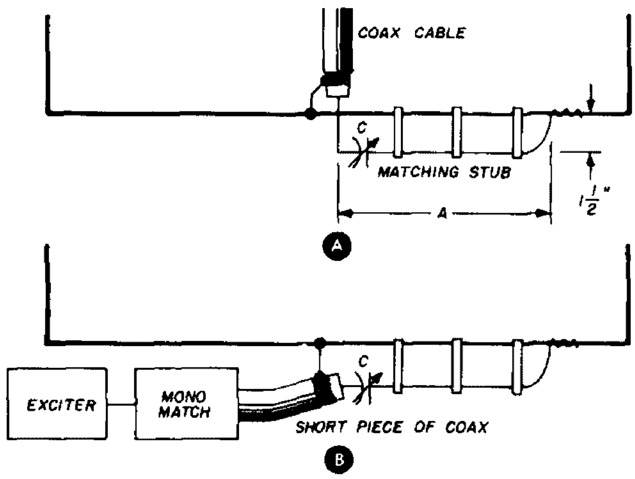



**fig. 4. The gamma match, and typical instrumentation for tuning.**

desire to operate on 20, 15, and 10 meters with high performance on all three bands, the quad definitely qualifies as a front runner. If, however, you wish to operate exclusively on one band, the monoband Yagi is probably first choice. Trap-element Yagis are in a class by themselves, and although they certainly have their place, their efficiency isn't as high as the quad or monoband Yagi.

## acknowledgements

I'd like to express my sincere appreciation to Karl Scharping, W6KWF, president of the Cubex Company, Altadena, California, for his assistance in the preparation of this article. Also, thanks are certainly due my wife, whose support and indulgence over the hours of experimenting made it all possible.

**ham radio**

# modular modulos

Mother and daughter printed-circuit boards are used to increase IC counter-circuit versatility

Ronald M. Vaceluke, W9SEK, 17W540 Hillcrest, Wood Dale, Illinois

The many articles on integrated circuits published during the past few years have helped these little plastic bugs gain wide acceptance by those who like to build their own equipment. The construction projects that didn't specify printed circuit boards usually turned out to be quite difficult because of the wiring required on the closely spaced pins of the ICs. IC sockets are really no help except when frequent replacement is necessary; besides, they're an added expense.

Let's face it—the PC board route is the only way to go when working with ICs. They are designed for PC board mounting, and any other mounting method just doesn't result in a neat, professional appearance. Of course, IC projects have been published specifying PC board construction. The only trouble here is that the circuits are restricted to the function for which they were originally designed.

## modular modulos

With these limitations in mind, I'd like to present what I call "modular modulos." The word "modular" is familiar to most readers, but "modulo" may not be, so a word of explanation is in order.

A modulo, or mod, is a circuit that

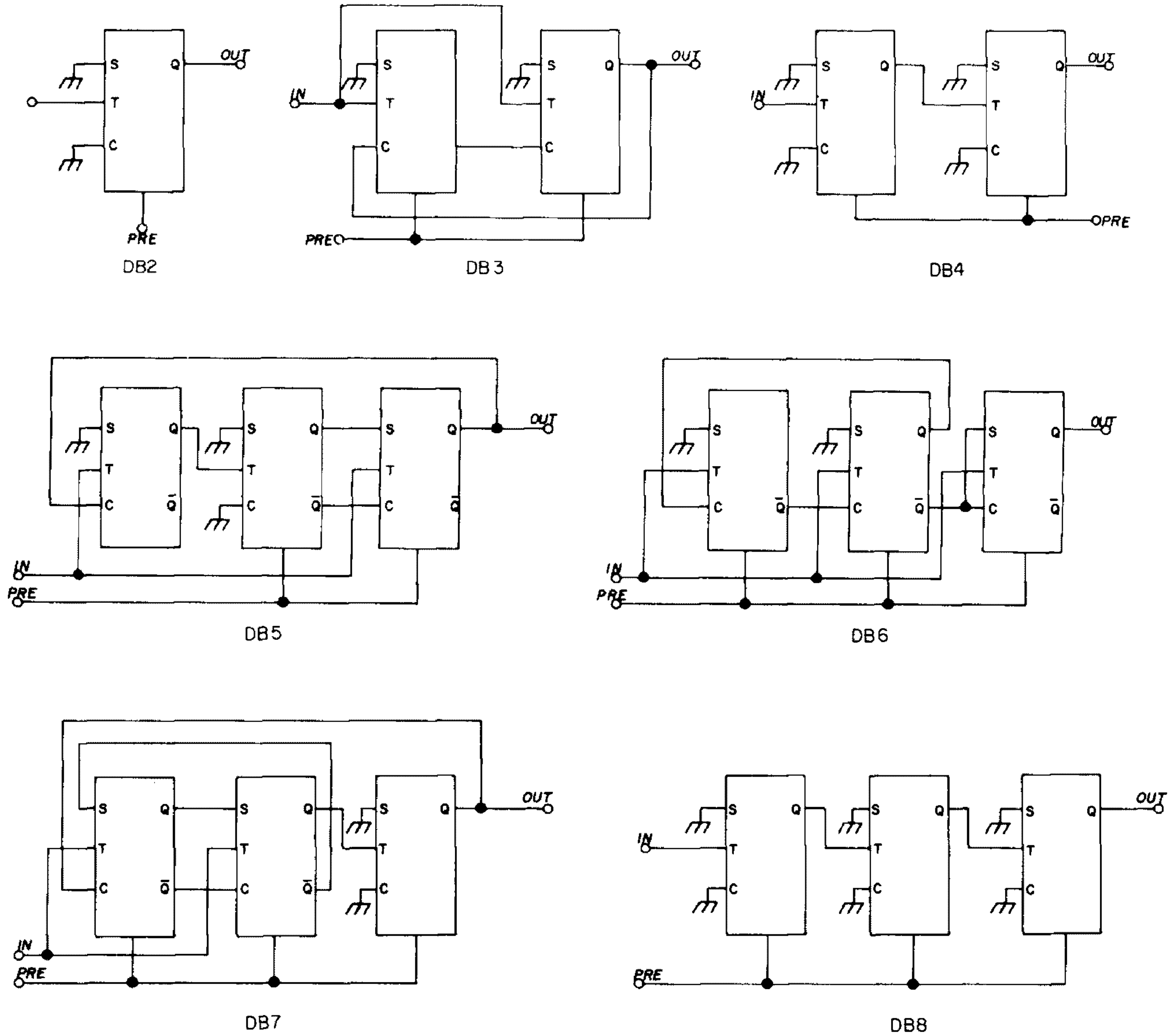


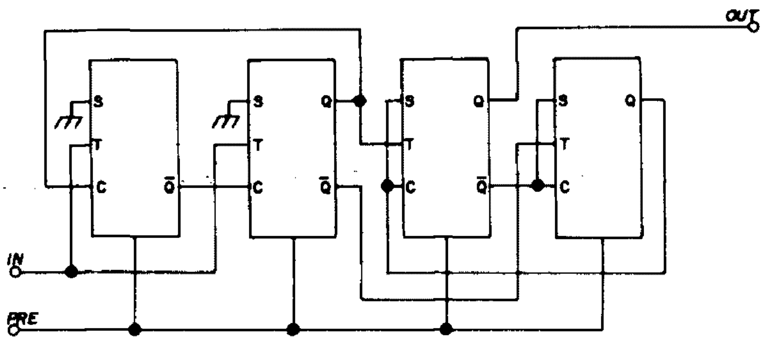

**fig. 1. Modulo circuits. Each circuit mounts on a daughter board, ten of which can be mounted on a mother board for a mix of counter functions. DB corresponds to "divide by;" the DB-3 circuit divides by 3.**

counts; or as some prefer, divides. A mod is designated by how many counts occur before an output is produced. A mod 7, for example, will count to 6 and the output will remain unchanged. When the 7th input occurs, the output will change to indicate that the count is complete.

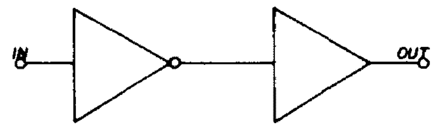
The resistor-transistor logic (RTL) ICs used here are edge-triggered by a sharp negative transition from logical 1 (approximately Vcc) to logical 0 (approximately ground). During a count cycle the modulo output will go from logical 0 (its reset condition) to logical 1.

The part of the count at which this action occurs will vary between modulos; that is, mod 5 will go high at some point in the count different from, say, mod 10. The transition from 0 to 1 will have no effect on following counters. However, when the count is complete and the output goes from 1 to 0, the following counter will be triggered.

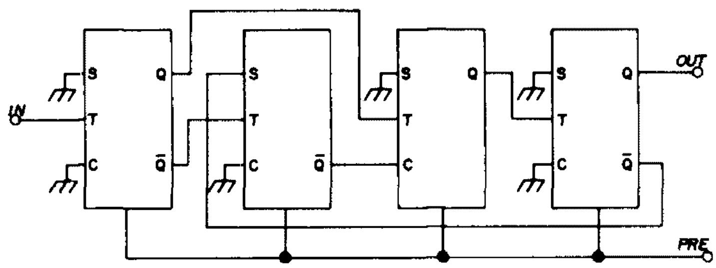
This output on a scope will appear as a train of square waves. If an accurate frequency source were connected to a number of modulos wired in tandem, a division of this frequency will give accurate submultiples from which time pulses can be obtained. These can be used to calibrate receivers, oscilloscope time bases, etc. The modulos described here were used for this purpose in the frequency and time standard shown in the photo.



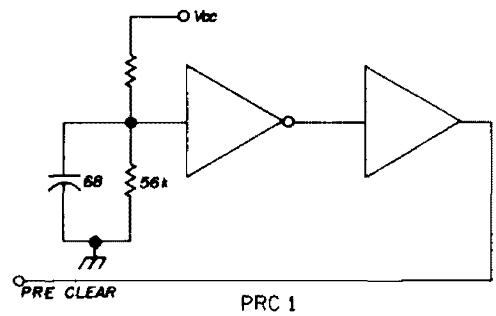

DB9



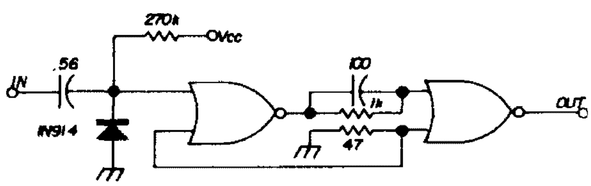

NIB 1



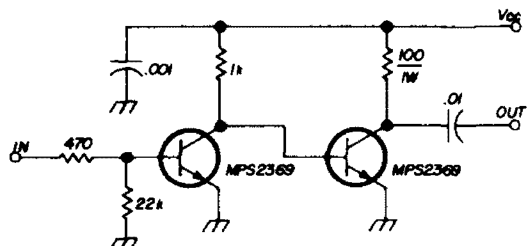

DB 10




PRC 1




ST 1




HA 1

The transistorized oscillator used in the standard is extremely stable. It should be remembered that no matter how many frequency divisions are made, the output is only as accurate as the input.

These modulos can also be used as a basis for a TV synch generator count-down chain (ATV'ers please note), or to obtain the gating pulses in a frequency counter. By dividing the 60-Hz power-line frequency by 60, 1-Hz pulses can be obtained for an electronic clock. Other uses will undoubtedly come to mind for these versatile circuits.

## construction

Each circuit (fig. 1) is mounted on a small PC board. These boards, called daughter boards, mount on a mother board (fig. 2) by Varicon connectors

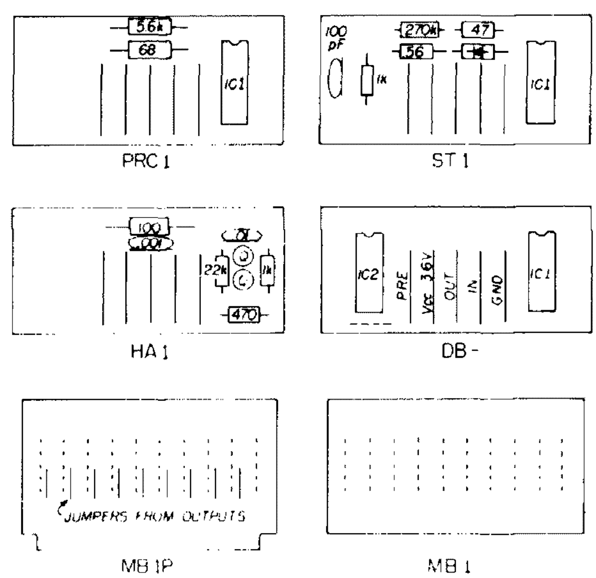



**fig. 2. Top of PC boards showing parts layout. Dashed jumper on DB board is required only for DB7 and DB9 circuits.**

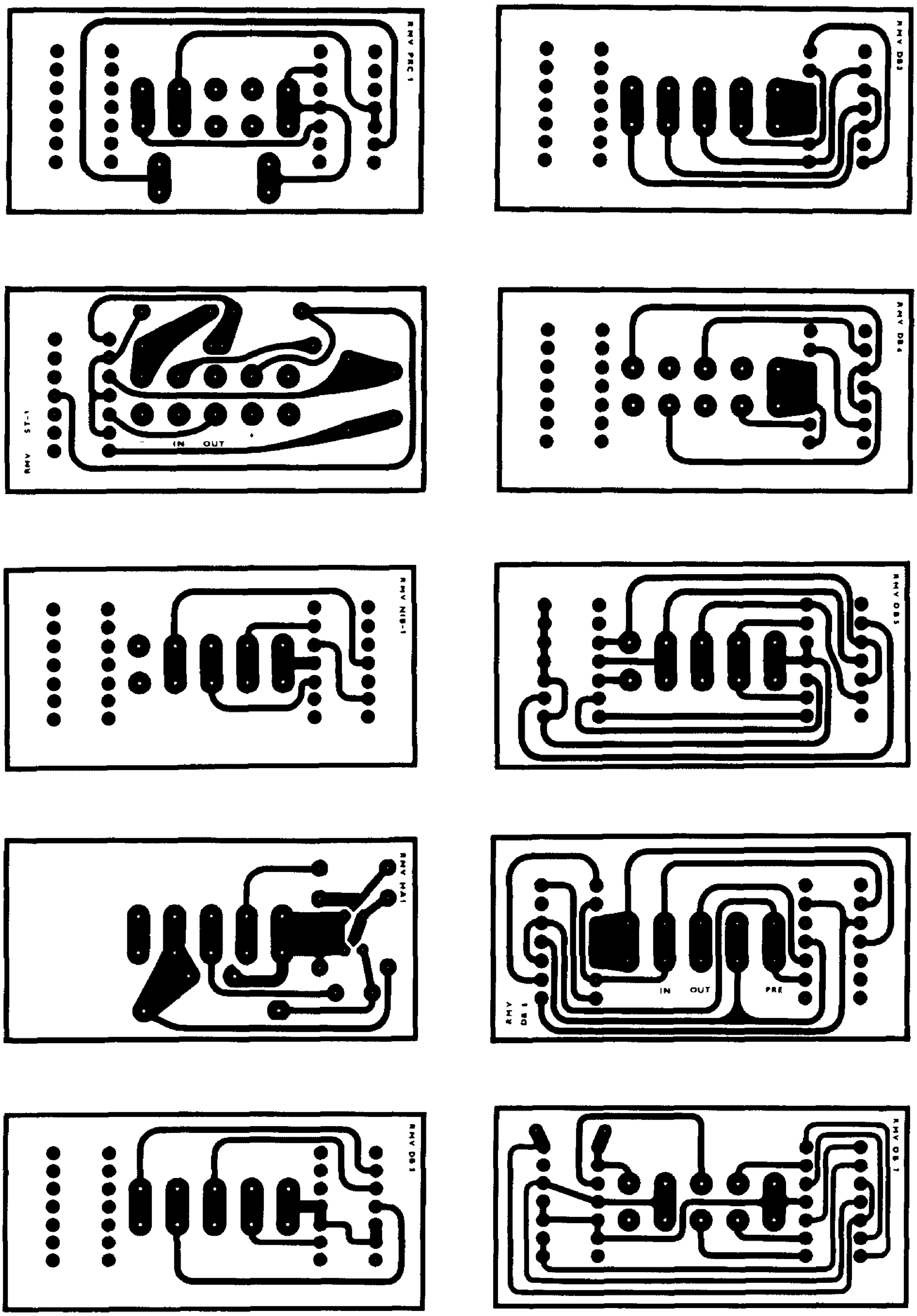
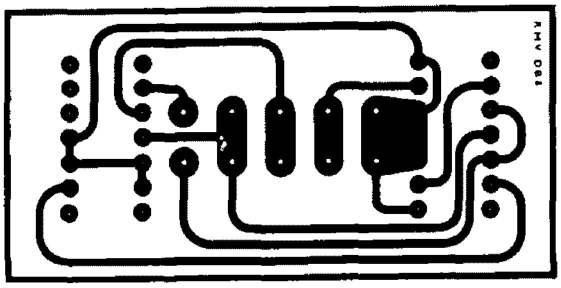
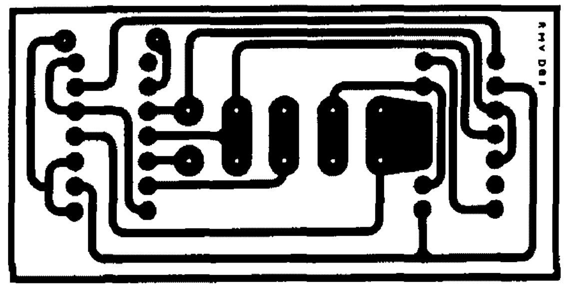
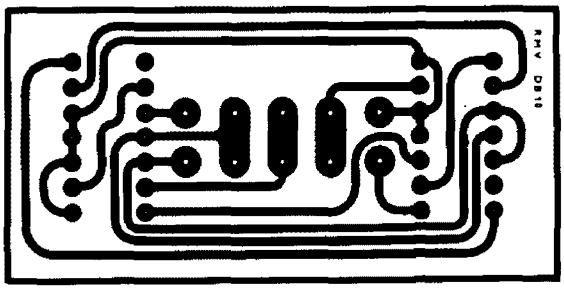
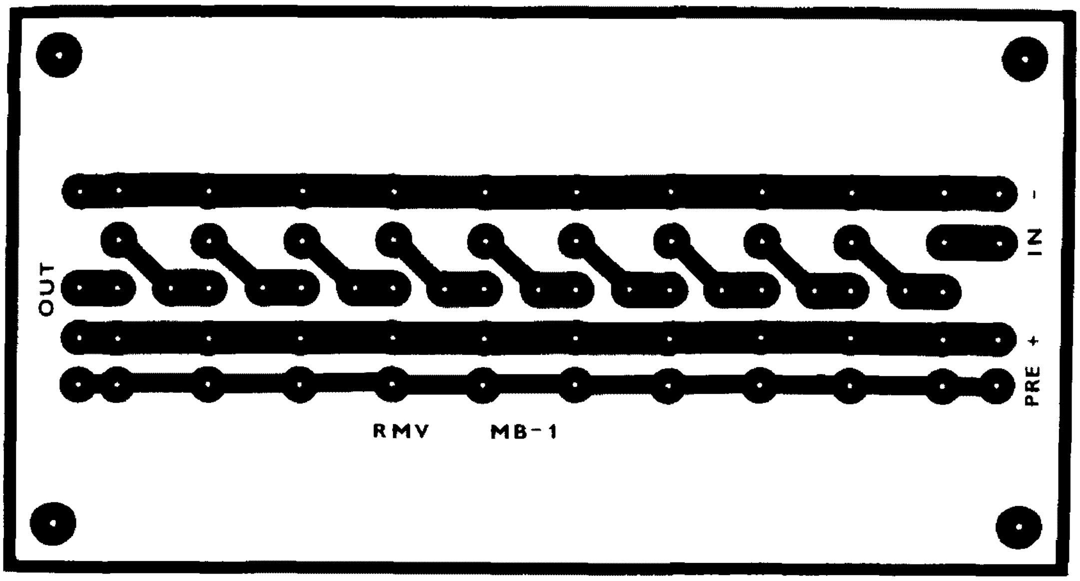



fig. 3. Etched side of mother and daughter boards.








OUT
−
IN
+
PRE
RMV MB-1

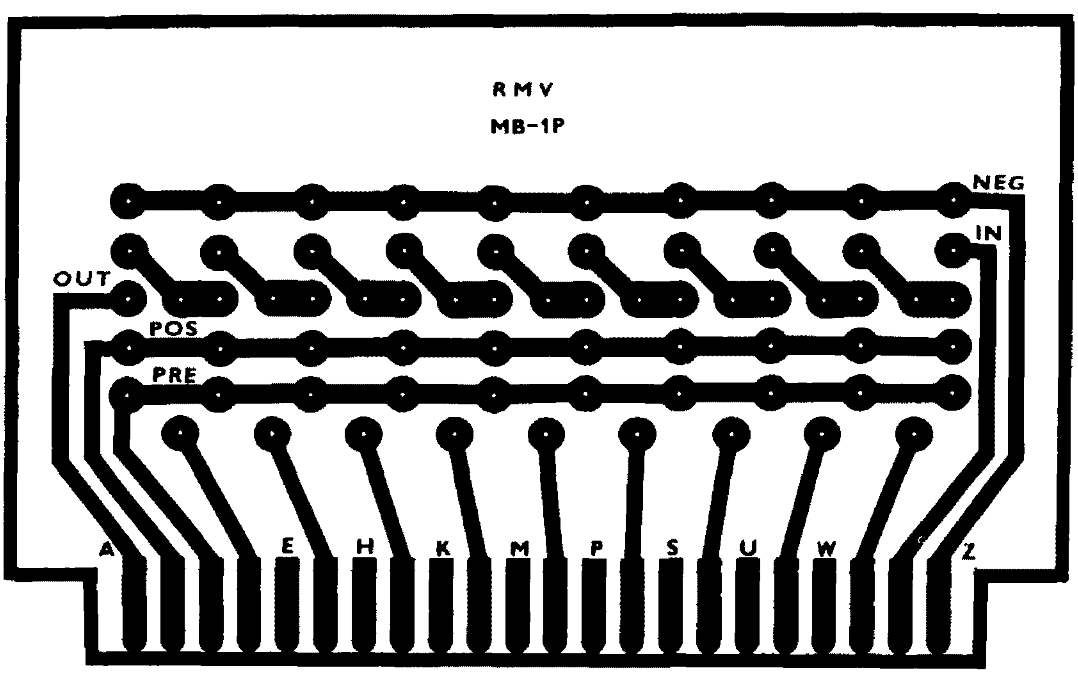

RMV
MB-1P
NEG
IN
OUT
POS
PRE
A
E
H
K
M
P
S
U
W
Z

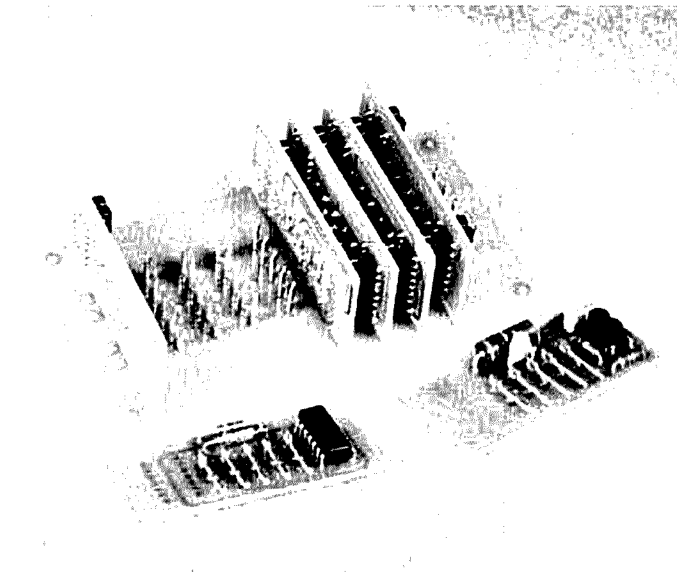
table 1. Loading factors for integrated circuits used in the counters and associated circuits.

| board | in | out | clear | IC1 | IC2 |
|---|---|---|---|---|---|
| DB-2 | 2 | 5 | 1 | MC726P | — |
| DB-3 | 4 | 4 | 2 | MC791P | — |
| DB-4 | 2 | 5 | 2 | MC791P | — |
| DB-5 | 4 | 4 | 3 | MC791P | MC791P |
| DB-6 | 6 | 5 | 3 | MC726P | MC790P |
| DB-7 | 4 | 5 | 3 | MC790P | MC726P |
| DB-8 | 2 | 5 | 3 | MC726P | MC790P |
| DB-9 | 4 | 5 | 4 | MC791P | MC790P |
| DB-10 | 2 | 5 | 4 | MC791P | MC790P |
| ST-1 | - | 5 | - | MC724P | — |
| PRC-1 | - | 25 | - | MC799P | — |
| NIB-1 | 2 | 25 | - | MC799P | — |

made by Elco. The daughter-board connector is no. 02-030-013-5200 and mates to the mother board with connector no. 02-030-113-6200. Thirty connectors are mounted on a plastic strip. They should be broken into segments of five connectors each. The PC etched board layout is shown in fig. 3; however, if it isn't convenient to etch your own boards, they may be purchased.* In any event, the use of a small-tipped, twenty-five-watt iron is recommended. It's best to mount the connectors first using the plastic strip to hold the connectors in alignment until soldered in place. After soldering, the plastic can be broken away.

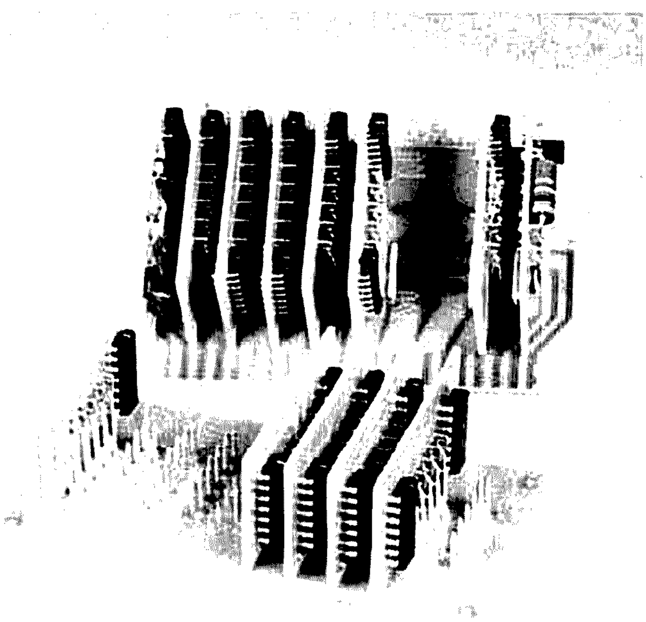
The mother boards are shown in two styles, and both will hold up to ten daughter boards. One mother board can be mounted on a chassis with stand-off insulators over a cutout. The other plugs into a socket, such as an Amphenol 143-022-01. This socket, as well as the Elco Varicons, can be purchased from Allied Radio in Chicago.

Mother board can be cut to shorter length if ten daughter boards aren't needed. Shown below is the pre-clear (left) and harmonic amplifier modules.



The output side of the daughter boards feeds the input to the next stage. The output connectors appear on a separate pad on the board as well. Leads can be connected to these points to give different ratios of division throughout the chain if desired. The plug-in-style mother board must have these points jumpered so they appear at the main connector. See

The two styles of mother boards. The plug-in type is used in the frequency and time standard. The other (below) can be mounted on a small chassis with outputs brought out to banana jacks.



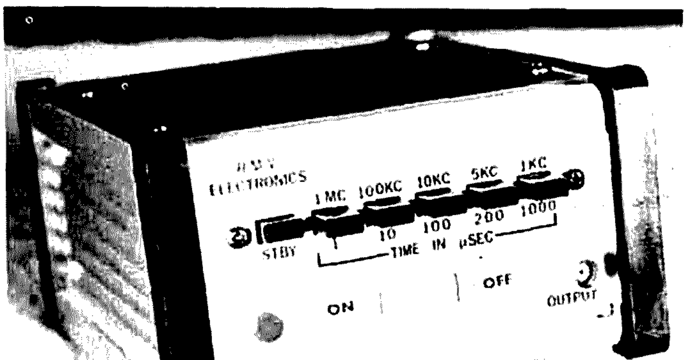
*Available from RMV Electronics, P. O. Box 283, Wood Dale, Ill. 60191. Daughter boards are $1.25 each postpaid in quantities of 1 to 5; $1.10 each in quantities of 6 and up. Mother boards are $2.00 each postpaid. All boards are tinned, drilled, and ready for components.

fig. 2 for the location of these jumpers.

## rules for use

The most important thing to observe is the loading factor, **table 1**. Do not let the input loading factor exceed the output loading factor (fan-out) of the preceding stage. A higher output loading factor than is needed is permissible. If the required input is greater than that available, use a buffer stage (NIB 1). Since the flip-flops are edge-triggered by a sharp drop from logical 1 to logical 0, frequencies below approximately 100 kHz must be squared-up with a Schmitt trigger (ST 1). This circuit can be used to the subaudible range: 7 Hz or less. If a high-harmonic output is desired (as in a frequency standard) the harmonic amplifier (HA 1) is *recommended*.

It is desirable to clear all stages before starting to count. This is done by first applying Vcc (3.6V) to the pre-clear line and then grounding this line when counting. This can be done manually or by the PRC 1 board, which will pre-clear the stages automatically as power is applied to the mother board. The voltage required by these circuits is 3.6 volts dc. A well-regulated and filter supply will provide the best service.




**Frequency and time standard using modular modulos.**

ham radio

# the ham notebook

## perking up your HW-17A

The Heath **HW-17A** certainly can qualify as one of the more popular 2-meter transceivers, judging from the many in use on this band. However, as with most equipment of this type, it's always possible to improve on the original design as new devices become available after the fact.

The broadband preamplifier described in this article will improve sensitivity, increase rejection of fm and tv images, and provide better agc action in the HW-17A. Improvement in agc is not to be passed over lightly, because it makes the S-meter much more useful and compensates for some of the loss in the automatic noise limiter circuit. An additional feature of the preamp modification is that no new holes need be drilled to accommodate the circuit.

Also included are changes to improve transmitter modulation and increase modulator bass response.

### preamp circuit

Nothing exotic or unproven is included in the preamplifier circuit **(fig. 1)**. This circuit (with a 3N140 mosfet) was used by many 2-meter enthusiasts in a converter described in an earlier issue of *ham radio.*[1] Although not shown in the schematic, the 40673 has internal protection for the gates. This eliminates the need for external diodes, which are found in many semiconductor front-end circuits to eliminate static burnout.

### preamp construction

A physical layout compatible with the HW-17A is shown in **fig. 3**. A thin copper U-bracket chassis bolts to PC board 85-205 in the HW-17A with a single machine screw, which is also used to secure the tuner to this board. All resistors are mounted on the opposite side of the bracket. Feedthrough capacitors are used for bypassing.

### modulation improvements

Not all HW-17A transceivers are fully modulated, despite several circuit modifications incorporated by the manufacturer. One change was very helpful in my case, although it did decrease my power output slightly. By increasing the value of the screen resistor in the final amplifier, input power is decreased, which allows higher modulation percentage and improved upward modulation. The exact value of the resistor is best determined by observing the signal on an oscilloscope. My experience indicated

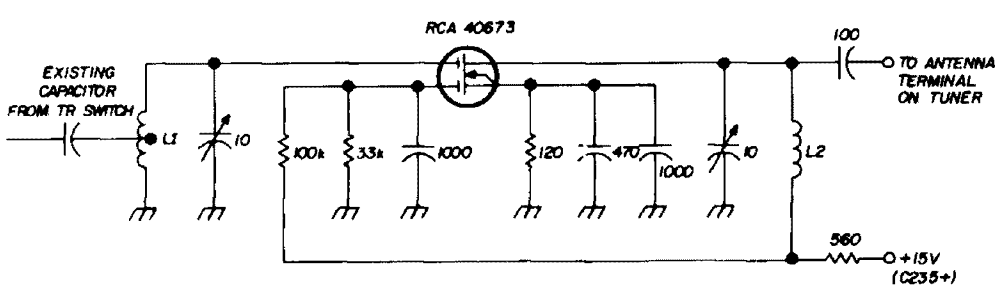



**fig. 1. Low-noise preamplifier using a tetrode mosfet for improving the Heath HW-17A receiver front-end response.**

that resistor R115 should be increased to 27k.

## modulator bass response

A final touch to improve the bass response of the HW-17A modulator will bring compliments on your audio quality. The emitter of Q11, a 2N3391, should be bypassed with a capacitor larger than the 2-μF unit in the original circuit. Unfortunately, the stage will motorboat with direct bypassing. My solution was to use a 56-ohm resistor in series with a 20μF capacitor, as shown in fig. 3.

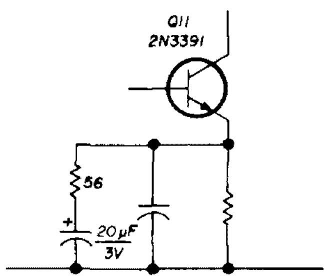



**fig. 2. Additional components for improving the HW-17A modulator bass response.**

## conclusion

The simple changes described above will more than repay you in increased performance and operating pleasure for the time and effort involved. When I made these modifications I was able to realize the best features of this fine transceiver. Now when DX is there, I work it!

### reference


1. D. Nelson, WB2EGZ, "High-Performance Mosfet Converter—the Two-Meter Winner," *ham radio,* August, 1968, pp. 22-29.


**Don Nelson, WB2EGZ**

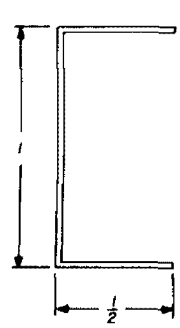



**fig. 3. Suggested parts layout for the preamp. Simple U-shaped copper chassis bolts to an existing pc board in the HW-17A.**

## removing IC's

I was accumulating parts for a project using ICs and needed a couple of JK flip-flops. I had some in my junk box, but the little rascals were firmly embedded in one of those bargain PC boards obtained from a surplus outlet. The board had been wave-soldered, which means the parts were there to stay put.

I didn't have a vacuum desoldering tool handy, so I used a trick known as wicking. I flattened a short piece of shield braid and held it against the pins of the IC. Applying the soldering iron to the braid causes it to absorb the solder, and the IC comes out easily. Everything must be clean: soldering iron, braid, and PC board.

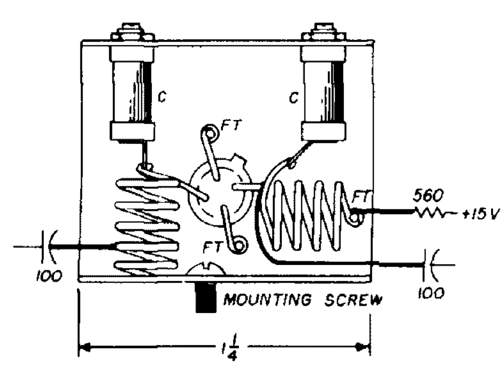
A vacuum desoldering tool is best if you have to remove many parts, but the wicking method is okay for one or two devices. I made a vacuum desoldering tool out of an old ear syringe, but it's slow and the tube tends to clog with solder. The best device is a spring-loaded tool designed for this purpose.

**Alf Wilson, W6NIF**

## two-meter converter

The converter shown in **fig. 4** receives on two frequencies. I use it with an auto radio to receive my vhf club stations on 145.35 MHz and an fm repeater on 146.94 MHz.

Most of the parts were salvaged from uhf television converters and transistorized bc receivers. Crystal frequencies were chosen to put the converter output into a clear spot in the broadcast band.



| | |
|---|---|
| L1 | 3 turns no. 22 enameled on ¼" form |
| L2 | 16 turns no. 24 enameled on ¼" form |
| L3 | Oscillator coil from transistor broadcast radio |
| RFC | 25 turns no. 26 enameled on R1 |

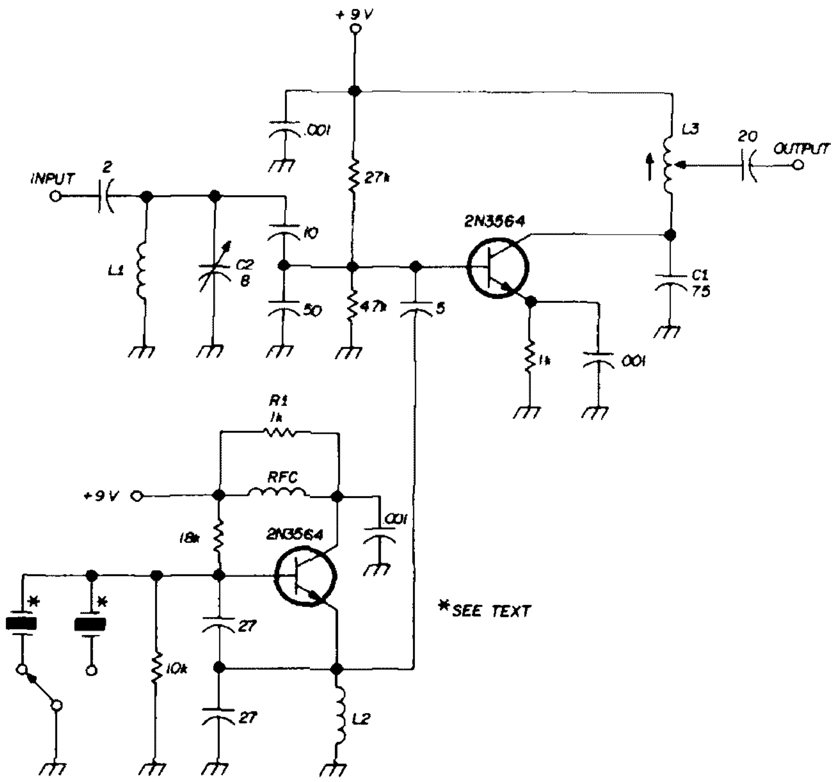



**fig. 4. Schematic for the two-signal vhf converter. Most of the components were salvaged from transistor radio receivers. Transistors are available from Poly-Paks or may be salvaged from late-model tv converters.**

## construction

The schematic is shown in **fig. 4**. The unit is built on a piece of circuit board and mounted in a small Bud box for complete shielding. The board should be insulated from the box if polarity is disregarded. (My box has a negative ground, so I didn't insulate the board.)

The output coil, L3, is an oscillator coil from an old transistor bc-band radio. The tap **(fig. 4)** was the emitter tap for the autodyne mixer. The unused winding was connected between the collector and i-f transformer. You may have to try different values of C1 to resonate your coil.

The antenna coil, L1, tunes rather broadly. It should be peaked between the two desired signals. Spread or squeeze L1 to make C2 tune to resonance.

## crystal selection

The crystals are third overtone types. The oscillator may be operated on either side of the input signal. For the low side, the crystal frequency will be

$$f_x = \frac{f_s - f_i}{3}$$

where

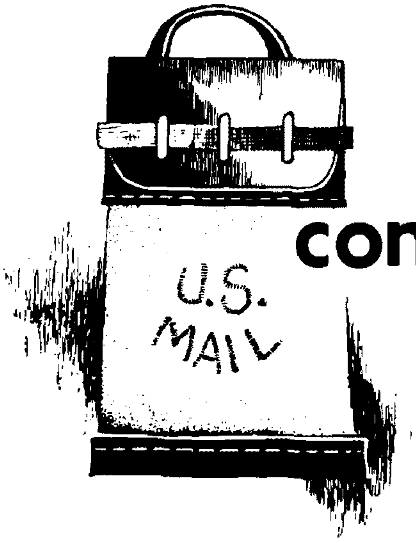
$f_x$ = crystal frequency
$f_s$ = signal frequency
$f_i$ = intermidiate frequency

To operate the oscillator on the high side of the input signal, merely add the signal frequency and intermediate frequency, then divide by 3.

I mounted a slide switch on the box to select the crystals. No signal feedthrough occurred since the converter and auto radio were well shielded.

This converter operates on 9 volts. If you use it in your car, I'd suggest you bypass the B+ line with a 25- to 50-$\mu$F capacitor. A dropping resistor and zener should be used to reduce the car's primary voltage.

**W. G. Eslick, K0VQY**




# comments

**Dear HR:**

During a recent QSO with a well-known amateur, the conversation turned to the quality and content of the general run of QSO's heard on the bands. What had started out to be just a casual contact turned into a lengthy discussion of the directions in which amateur radio is headed.

It seems that, more and more the prevailing idea is to contact a fellow, exchange signal reports and equipment types, and then cut and run to the next one. This situation is aggravated by the almost continuous stream of contests sponsored by one group or another, the purpose of which seem to be the maximum number of contacts in the minimum length of time.

I will be the first to acknowledge that amateur radio is a hobby open to all types and persuasions. Those whose main interest lies in the compilation of an impressive list of contacts and a wall full of certificates attesting to their ability in multiplication, are and should be free to pursue their aims. However, must we all be forced into this type of operation? Listen in any evening on the DX bands and count the number of genuine QSO's going on.

Some amateurs, in self-defense, have grouped together into small regional groups and don't even attempt, nor in some cases welcome, outside contact. The foreign amateurs seem to be getting more reluctant to answer calls, and I suspect it is for precisely this reason.

If we wish to retain our bands, perhaps it is not too early to examine the purposes for which they were granted in the first place and the utilization that is being practiced today. A QSL card used to be something a fellow sent to another station to acknowledge a pleasant contact, not something to confirm a statistic.

Perhaps we need to call *The Old Man* back from his well-earned rest to keep watch on us. His ever-ready Wouff-hong and threat of the Rettysnitch struck fear into the hearts of those who were guilty of "rotten" practices.

This station is open anytime I'm on the air to contacts with others that want to talk about something more meaningful than, "ur RST 5NN eqpt hr Foghorn II tnx qso diddley-bump-de-bump."

**Jim Crouch, K4BRR**
**Jacksonville, Florida**

## mnemonics

**Dear HR:**

Here's a mnemonic for you which I learned from Dr. Nolde, my first Electrical Engineering Professor at Berkeley in 1956.

ELI the ICE man.

ELI voltage leads (comes before current in an inductor

ICE current leads voltage in a capacitor.

Thanks for the article—I can use some of the ideas with my students.

**Les Hamilton, PH.D., K6JVE/3**
**Prince Georges, Maryland**

# new products

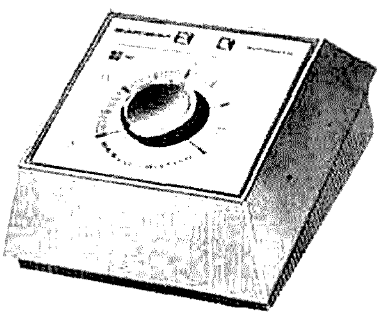
## electronic keyers


Curtis Electro Devices has announced two new integrated-circuit keyers that incorporate iambic-mode squeeze keying into the basic *Electronic Fist*. These new designs provide iambic-mode character generation (alternate dots and dashes) when both keyer levers are closed. The EK-39 Iambic Deluxe model includes dot memory, self-completing dots, dashes and spaces plus instant-start circuits that ensure easy and accurate character generation. The *weight* control, which is completely independent of the *speed* control can be used to lengthen characters according to individual preference.

A built-in monitor oscillator and speaker include an external volume control and internal pitch control. Provision is made for solid-state or relay switching. The unit will key grid-block rigs up to −150 volts at 50 mA with output transistor provided. A back-panel jack allows connection of a straight key. Three popular color combinations allow you to match your other station equipment. All cables and connectors are provided.

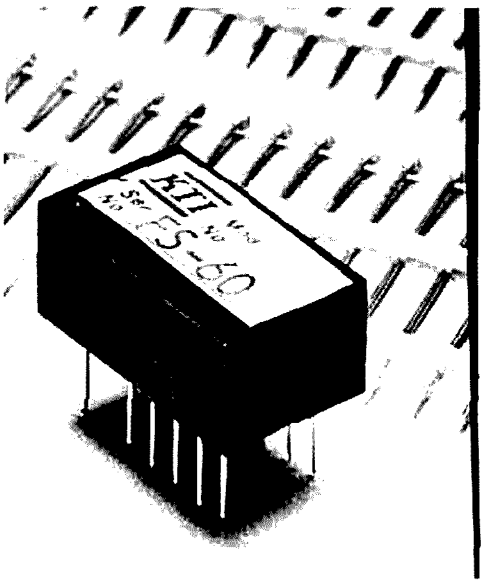
The EK-39 is available as a kit for $87.95, or factory wired and tested for $97.95. The model EK-38, which has all the features of the EK-39 except the *weight* control, is $72.95 as a kit, $82.95 wired and tested. A reed relay option is $4.00 additional. For more details, write to Curtis Electro Devices, Box 4090, Mountain View, California 94040.

## antenna rotator



Radio Shack's new antenna rotator, the *Servo-Rotor,* although designed primarily for television antennas, will handle a three-element 20-meter beam according to the manufacturer. The *Servo-Rotor* uses a modern solid-state amplifier/control circuit that compares actual antenna direction with the heading at the control box, and corrects any errors. Pointing accuracy is claimed to be ±0.5%. Special clutch action overcomes the effects of ice and wind, and a built-in brake eliminates overshoot. Indicator lights show turning direction. Heavy-duty V-shaped clamps attach the rotator to masts from 1½ to 2-1/8 inches in diameter. $39.95 from your local Radio Shack store.

## active filter




Kinetic Technology, Inc., has introduced a new hybrid integrated-circuit active filter that features many interesting characteristics. The low-power FS-60 filters require only 0.3 mW of power at ±2 volts, making it particularly suitable for battery-operated equipment. The filter works in the frequency range of dc to 10 kHz and has multi-loop negative feedback for high stability. Q range is from 0.1 to 500. The voltage gain of the unit is adjustable to 40 dB. Bandpass, high-pass and low-pass outputs are available simultaneously. The center frequency and Q can be tuned by adding external resistors.

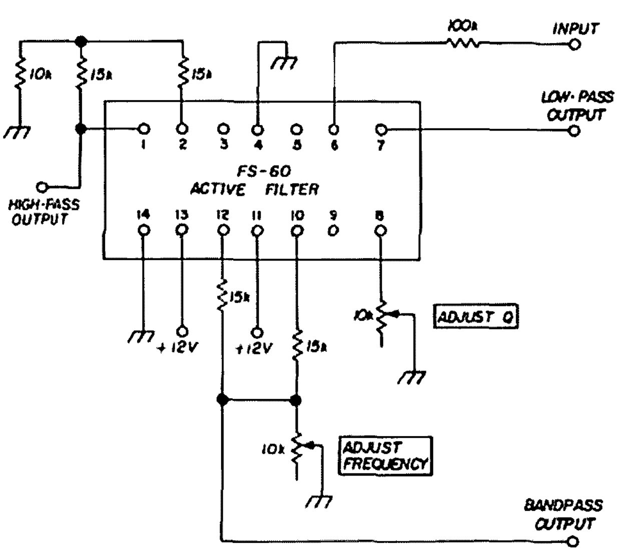
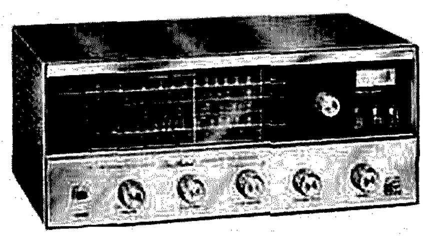



**fig. 1. Use a series resistor (100 to 1000 ohms) in the chosen output line to isolate the FS-60 possible capacitive or inductive loads. Device will not drive less than 10k load without reduced output swing.**

A typical amateur application of the FS-60 is shown in fig. 1. This circuit tunes from 200 to 800 Hz and makes a good outboard CW filter. Practically all the active moonbouncers are using this type filter and report that they have nothing better. It's even been rumored that Sam Harris, W1FZJ/KP4, has his chained to the bench! In the near future we expect to have more applications information on this interesting new device in *ham radio.*

For complete specifications and pricing information, write to Kinetic Technology, Inc., 3393 De La Cruz Boulevard, Santa Clara, California 95051.

## general-coverage receiver



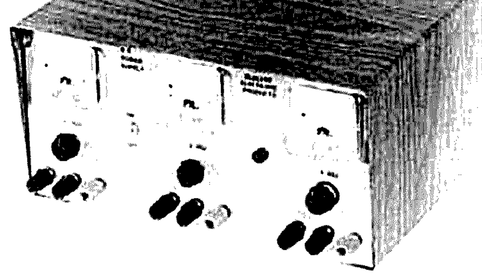
Radio Shack has introduced a new solid-state general-coverage receiver that tunes the entire range from 535 kHz to 30 MHz. The new DX-120 receiver features an fet in the front end to minimize cross-modulation from strong local stations, but still provides good sensitivity for weak signals. Four color-coded tuning scales and a logging scale provide easy readability; fine tuning is performed with a separate bandspread control. An S-meter, variable bfo and automatic noise limiter are provided; a built-in regulated power supply stabilizes receiver power supply-voltage drift and battery aging. It can be plugged into your local 117 Vac line, or used with 12 Vdc battery power for portable operation. Front panel design is a modern black with brushed-aluminum trim; the wrap-around cabinet is steel. The DX-120 can be operated with it's own built-in speaker, or with external headphones. $69.95 from your local Radio Shack store.

## how to use test instruments in electronics servicing

This new book by Fred Shunaman, former editor of *Radio-Electronics* is different from the usual books of this type. Instead of a series of "how-it-works" descriptions, this manual describes specific tests and troubleshooting techniques for the electronic experimenter. The first two chapters explain new ways to use an oscilloscope in tv troubleshooting, the next two chapters present tricks that can be done with vom's and vtvm's. Signal generators are covered in the next chapter, and the text describes signal-injection troubleshooting, alignment techniques, and how to measure inductance and capacitance with a signal generator.

Other chapters cover capacitor checks, test probes, tube and transistor checks, crt testers and rejuvenators, cathode-current measurements, field-strength measurements, signal tracing, impedance and frequency measurements, and maintaining and calibrating test equipment. One whole chapter is devoted to color-tv test gear, revealing a number of important pointers (and pitfalls) facing the technician. 256 pages. $4.95 from your local electronics distributor, or write to TAB Books, Blue Ridge Summit, Pennsylvania 17214.

## variable power supplies



Blulyne Electronics Corporation has announced a new line of economical, continuously-variable low-voltage power supplies that are ideal for semiconductor work. The units feature excellent regula-

tion and ripple characteristics, and are short-circuit proof. Single, dual and triple units are available.

The single unit (PS-61C) provides ±15 Vdc output with a load current of 700 mA (usable up to 1 ampere over 10 Vdc). Regulation is 0.0005% Vdc per mA load current; ripple is less than 0.005 at full load. Price: $49.95.

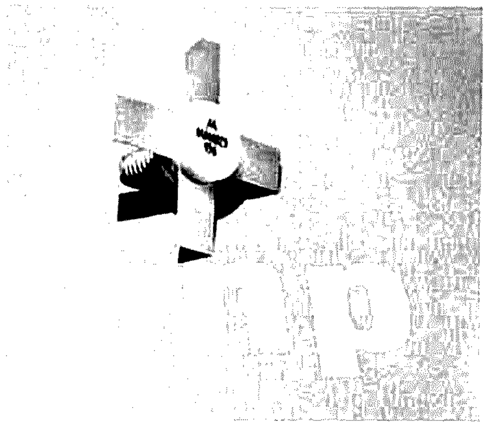
The double unit (PS-62C) is ideal for working with operational amplifiers and has the same electrical specifications as the PS-61C at each output. Price is $74.95. The triple unit (PS-63C) is three completely independent power supplies that may be used in any combination. Each output has the same electrical specification as the PS-61C. Price is $99.95.

All power supplies are furnished in modern walnut-finished cabinets. The grounding plug, single-switch operation, chassis-ground terminals and isolation from the ac line provide complete safety. For more information, write to Blulyne Electronics Corporation, 3 Sand Springs Road, Williamstown, Massachusetts 01267.

## multi-antenna coupler

The new multi-antenna coupler from Antennalabs is designed to couple several different antennas to a single transmission line. The coupler separates incoming signals according to frequency, and routes signals in each band to a separate terminal. It can be used with any 52-ohm unbalanced load (72-ohm models available), and may be located any convenient distance from the antenna since feedline length is not critical. The coupler is encapsulated in a durable weatherproof epoxy housing; power rating is 2000 watts, insertion loss 0.5 dB maximum, isolation between terminals is 20 dB and vswr is less than 1.5:1 on all bands. Models are available for all amateur bands from 3.5 to 144 MHz. Prices range from $24.95 to $34.95 depending on frequency. For more information write to, Antennalabs, Post Office Box 458, Ocean Bluff, Massachusetts 02065.

# pnp rf power transistors



Motorola Semiconductor has added four new transistors to their line of pnp rf power devices. With the addition of the new transistors, types MM4020 through MM4023, the line now covers the power range from 0.5 to 40 watts at 175 MHz. Each of the transistors features balanced-emitter construction for maximum safe operating area, isothermal design for flat power output vs temperature, and low inductance stripline packaging.

Balanced-emitter construction uses thin-film nichrome resistors in series with each of the multiple emitters to distribute power evenly throughout the semiconductor chip. Isothermal design results in exceptional output power stability with changing operating temperature. This insures even generation and flow of heat from the chip so that the power slump usually encountered in an rf power device operating near its maximum frequency is nearly absent. The stripline package provides minimum series lead inductance, and when combined with the inherently higher input impedance of pnp transistors, simplifies the design of interstage matching circuits; it also results in higher gain-bandwidth products.

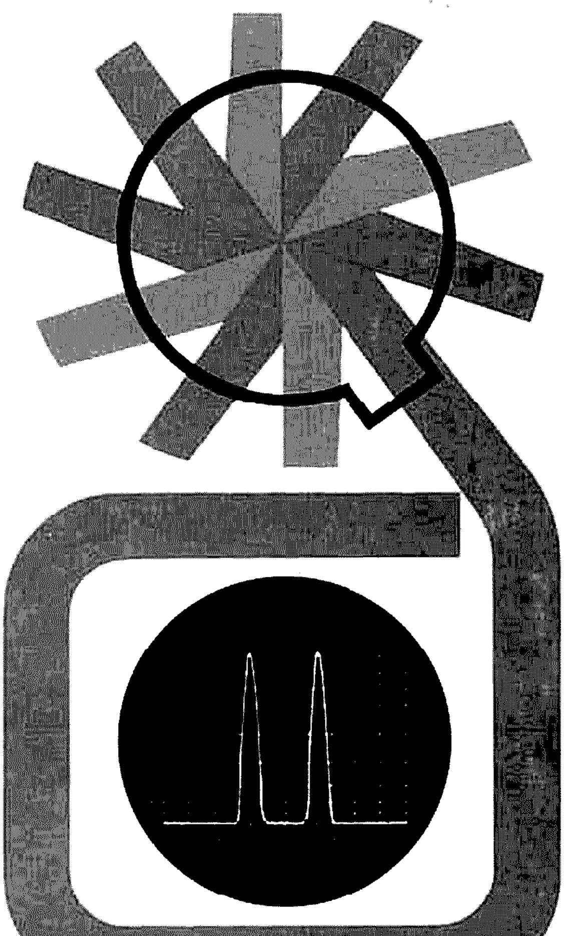
At 175 MHz, the MM4020 provides 3.5 watts output with 11.5-dB gain; the MM4021 provides 15 watts (7.0-dB gain), the MM4022 provides 25 watts (5.5-dB gain), and the MM4023 provides 40.0 watts at 4.5-dB gain. For complete specifications, write to Technical Information Center, Motorola Semiconductors Products, Inc., Box 20924, Phoenix, Arizona 85036.

75 cents

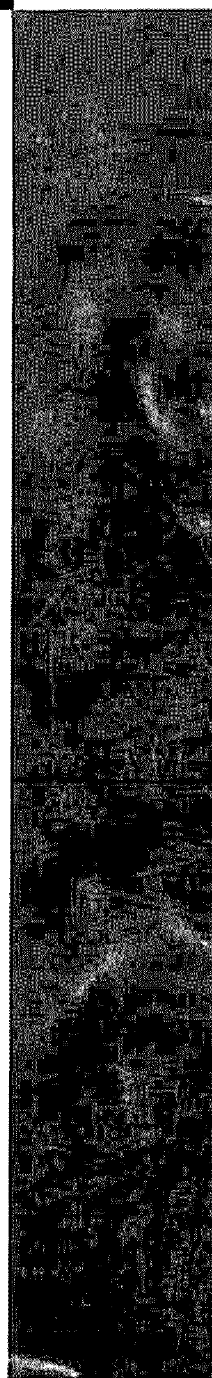
focus on communications technology . . .

# ham radio magazine

SEPTEMBER 1970



an
## INTEGRATED-CIRCUIT BALANCED MODULATOR



## this month

september 1970
volume 3, number 9



**offices**

Greenville, New Hampshire 03048
Telephone: 603-878-1441

**ham radio** magazine is
published monthly by
Communications Technology Inc.
Greenville, New Hampshire 03048

Subscription rates, world wide:
one year, *$6.00, three years, $12.00.*
Second class postage
paid at Greenville, N. H. 03048
and at additional mailing offices

Foreign subscription agents:
United Kingdom:
Radio Society of Great Britain,
35 Doughty Street, London WC1, England.

All European countries:
Eskil Persson, SM5CJP, Frotunagrand 1,
19400 Upplands Vasby, Sweden.

African continent:
Holland Radio, 143 Greenway,
Greenside, Johannesburg,
Republic of South Africa


Printed by Wellesley Press, Inc.
*Wellesley, Massachusetts 02181, U.S.A.*

**ham radio** is available to the blind
and physically handicapped on magnetic tape
from Science for the Blind,
221 Rock Hill Road, Bala Cynwyd,
Pennsylvania 19440.
Microfilm copies of current
and back issues are available
from University Microfilms,
Ann Arbor, Michigan 48103.

Postmaster: Please send form 3579 to
**ham radio** *magazine*, Greenville,
New Hampshire
03048

**ham radio** *magazine*

# contents

# a second look by jim fisk

**Microwave acoustics** is a new technology that promises to revolutionize communications equipment design in the future. Recent laboratory experiments with microwave acoustics have resulted in amplifiers, oscillators, resonators, signal couplers and delay lines. Nor is this a scientific curiosity confined to the inner-sanctum of the lab—Zenith is working on a 40-MHz acoustic-wave bandpass filter to replace the tuned circuits in color television sets.

The word "microwave" has traditionally been used to describe work in that part of the spectrum where wavelengths are defined in terms of centimeters. However, in "microwave acoustics" the term "micro" is associated with the micron wavelength of an acoustic wave on the surface of a crystal—at 30 MHz, for example, a 100-micron wavelength is possible because acoustic waves travel 100,000 times slower than electro-magnetic radiation.

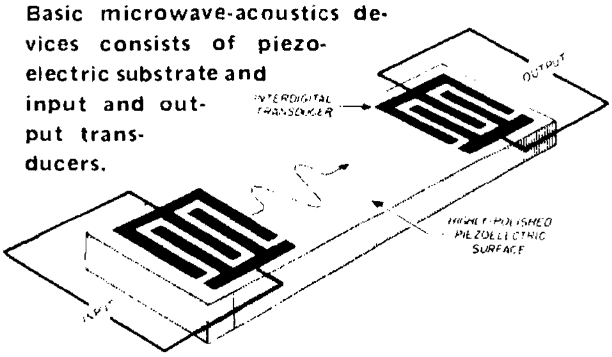
At the heart of all acoustic-wave devices is the delay line shown below. It consists of a piezoelectric substrate, such as quartz, and input and output transducers. The input transducer is basically a transmitting antenna that converts the incoming electrical signal into an acoustic wave on the surface of the substrate. When the acoustic wave reaches the opposite end of the substrate it is converted back into an electrical signal by the output transducer. The wave propagates rather slowly across the surface of the substrate, and by changing the spacing between transducers signal delay can be controlled.

Transducer design is extremely important because mass and shape affect efficiency, and size affects bandwidth. The interdigital structure shown in the drawing consists of two separate arrays of metal electrodes which resemble interlaced fingers. By changing the number of fingers the bandwidth can be tailored to circuit requirements. Thus, the basic acoustic-wave delay line becomes a resonator that may be used in place of tuned LC circuits.

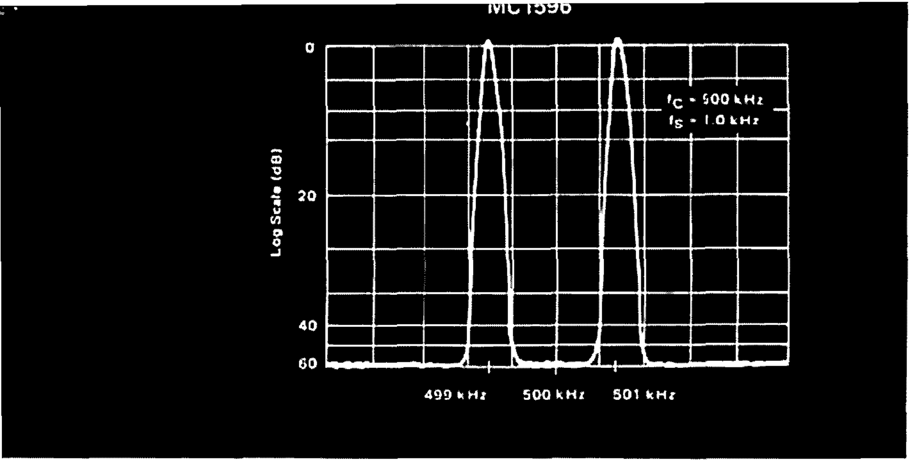
Engineers have recently come up with an acoustic-wave amplifier that uses a semiconductor structure spaced a short distance away from the substrate. Since the electric field associated with the propagation of the surface wave extends out of the surface of the substrate it interacts with the electrons in the semiconductor. If the dc supply voltage is low, the surface wave is attenuated because energy flows from it to the slower moving electrons in the semiconductor. However, as the supply voltage is increased, the electrons speed up, and when their speed exceeds that of the surface wave, gain results. Laboratory-built amplifiers using this system have netted gains on the order of 30 dB.

Although it will be some time before these new devices will be available for amateur use, they lend themselves to batch processing. And batch-processed components mean big savings after the design becomes standard—look at the proliferation of low-cost batch-processed ICs currently on the market.

**Jim Fisk, W1DTY**
**editor**




**Basic microwave-acoustics devices consists of piezoelectric substrate and input and output transducers.**



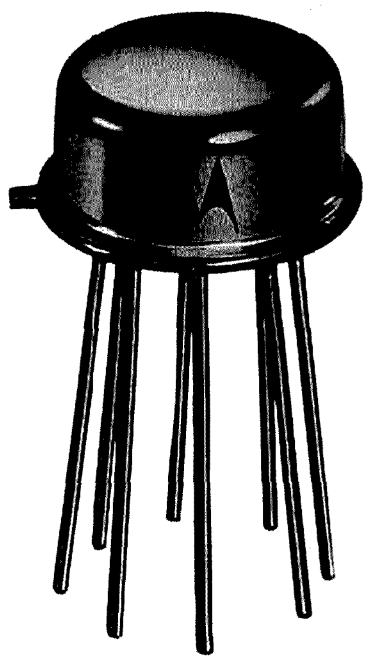

# an integrated-circuit balanced modulator

Roy Hejhall, K7QWR, P. O. Box 3265, Scottsdale, Arizona 85257

Although designed for balanced-modulator service, the Motorola MC1596G is readily adaptable to many other circuits for amateur use

Integrated circuits are being designed that can perform more and more of the circuit functions in amateur communications equipment. This article describes one of these new circuits, the Motorola MC1596G balanced modulator. Included are circuits showing the MC1596G as a balanced modulator and several other applications including an a-m modulator, a-m detector, a product detector, a mixer, and a frequency doubler.

**The Motorola MC1596 integrated-circuit balanced modulator contains 8 transistors, 3 resistors and 1 diode.**



The MC1596G is available now. However, a less-expensive version of the device with a slightly lower carrier-suppression specification and a limited operating-temperature range will be available soon under the type number MC1496G. The MC1496G will still provide very adequate performance for many amateur applications, and the circuits and information in this article apply to the MC1496G as well.

MC1596 performs this multiplication is beyond the scope of this article, and interested readers are referred to the references. However, to put the MC1596G to work it's helpful to have a basic knowledge of the output signal characteristics. Therefore, a brief discussion of the multiplication process with two ac signals follows.

Let's assume we have two sine-wave input signals called A and B at frequencies $f_A$ and $f_B$ respectively. And suppose we have a device that multiplies signal A times signal B and produces a third signal, C, which is the product of A and B. A device that performs this task is the MC1596G, and signal C will then have the following characteristics:

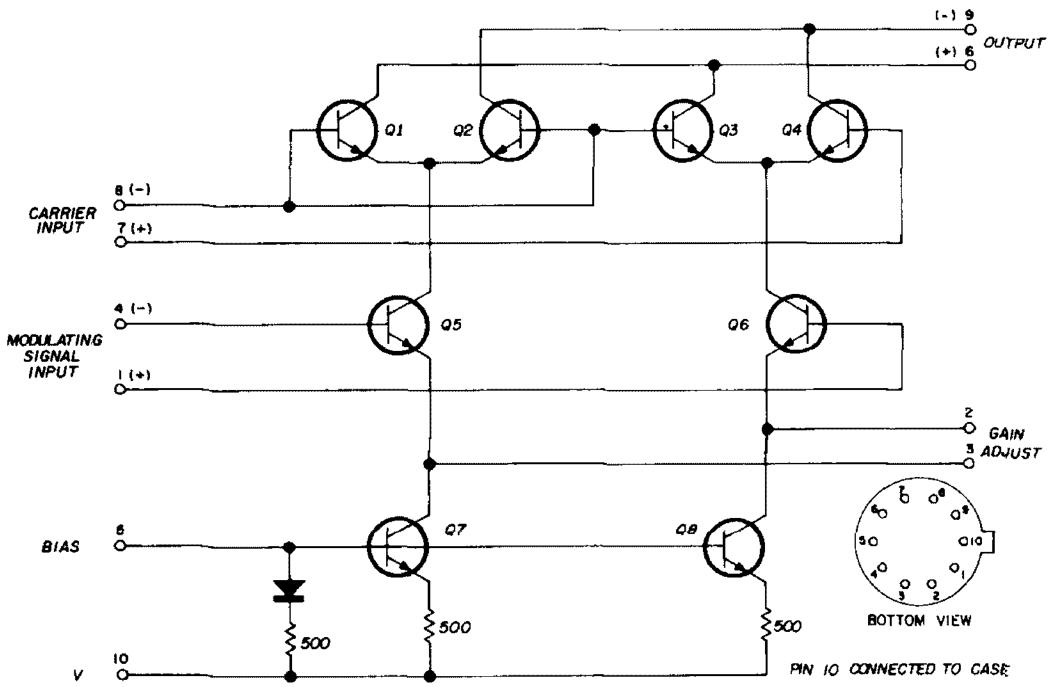



**fig. 1. Internal circuit arrangement of the Motorola MC1596G IC.**

## description of the MC1596G

**Fig. 1** is a schematic of the MC1596G. It consists of a dual differential amplifier driven by a standard differential amplifier. Transistors Q1 through Q4 make up the dual (upper) differential amplifier, while transistors Q5 and Q6 form the standard (lower) differential amplifier. Transistors Q7 and Q8 are constant-current sources for the lower differential amplifier.

The MC1596G has terminals for two input signals and one output signal. In operation, the circuit produces an output signal that is the product of the two input signals. A detailed discussion of how the

1. The amplitude of signal C will be the product of the amplitudes of signals A and B.

2. Signal C will contain two (and only two) frequency components, $(f_A + f_B)$ and $(f_A - f_B)$.

Note there is no output at either of the input signal *frequencies*, $f_A$ and $f_B$.

An example may be helpful at this point. Suppose we apply two input signals, one at a frequency of 1 MHz and the second at 4 MHz. The output signal will then contain frequency components at 3 and 5 MHz. In other words the output will be two separate, single-frequency sine-wave signals, one at 3 and one at 5 MHz. There will be no output at 1 and 4 MHz.

The signal amplitudes need a little further clarification at this point. The MC1596G has no built-in output load resistors; they must be added externally to develop an output-signal voltage from the output current. The magnitude of this output-voltage depends on the value of these outboard load resistors. Therefore, varying the load resistors will change the gain of the MC1596G. This means that the output signal will be the product of the input signal amplitudes times some constant, which is a function of the external load resistors and other circuit adjustments that affect gain. These gain-related items are described next.

## gain considerations

In addition to the load resistances, two other parameters that affect gain, or amplification, from the modulating-signal input to output are the magnitude of the resistance between pins 2 and 3 and the dc-bias currents. For all applications shown here, a bias current of 1 mA in each transistor of the lower differential amplifier, Q5 and Q6, has been used. This is generally recommended for most applications.

The resistance between pins 2 and 3 can be readily tailored to the needs of a particular circuit. Increasing this resistance decreases gain but increases the signal-handling capability of the IC. This means that output-signal amplitude will be decreased, but a higher-amplitude input signal can be handled without distortion. Decreasing this resistance has the opposite effect. The resistance between pins 2 and 3 can be anything between zero and several thousand ohms, depending on the optimum desired combination of gain and signal-handling capability.

Now let's see how the MC1596G can be put to work.

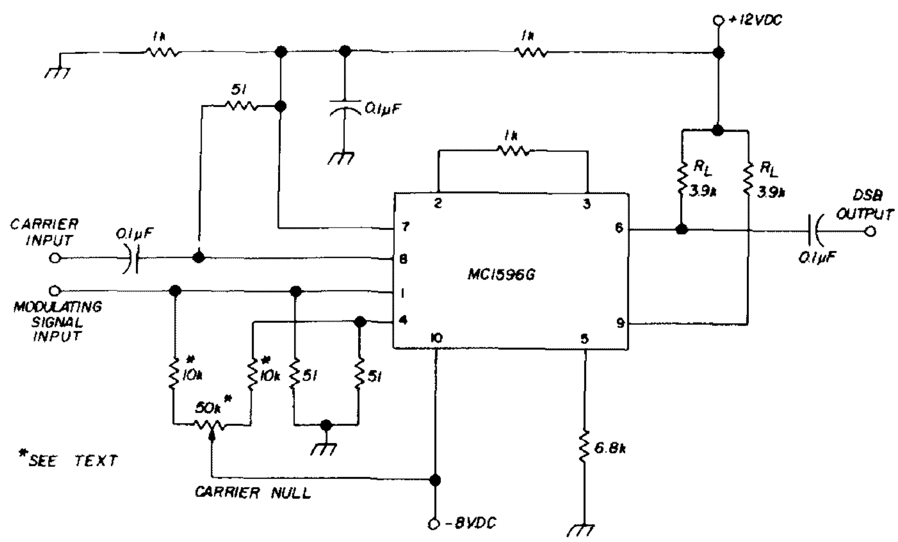



**fig. 2. Balanced modulator circuit. For use as an a-m modulator, merely change the two 10k resistors to 750 ohms. The pot can be adjusted for carrier null (ssb) or carrier insertion (a-m).**

## balanced modulator

**Fig. 2** shows the MC1596G as a balanced modulator. This circuit has the following advantages over a conventional diode ring balanced modulator often found in amateur equipment:

**1.** Circuit simplicity. No transformers are required; only resistors and capacitors. Only a single carrier-null adjustment is used, while diode ring modulators often have two null adjustments. Further, the MC1596G carrier-null adjustment is in the dc portion of the circuit. This means the carrier-null potentiometer need not be located physically near the remainder of the balanced-modulator circuit—it could be located on the rear panel of the equipment or at any remote location.

**2.** Greater carrier suppression. This balanced modulator will provide typically 65 dB and 50 dB carrier suppression at 500 kHz and 9 MHz respectively.

**3.** Broadband operation. The basic circuit requires no modifications for carrier frequencies from audio to 100 MHz.

The balanced modulator shown in **fig. 2** also has an extremely clean double-sideband output signal and a low-carrier-oscillator power requirement. Spurious sidebands are down 55 dB or more, and the carrier oscillator must deliver only 0.072 milliwatt to the modulator.

Construction and operation are simple. Reasonable care is needed to isolate output and carrier input. The modulator has such excellent carrier suppression that if care is not paid to circuit input-output isolation, there may be more carrier output signal passed through circuit stray capacitance than through the IC itself.

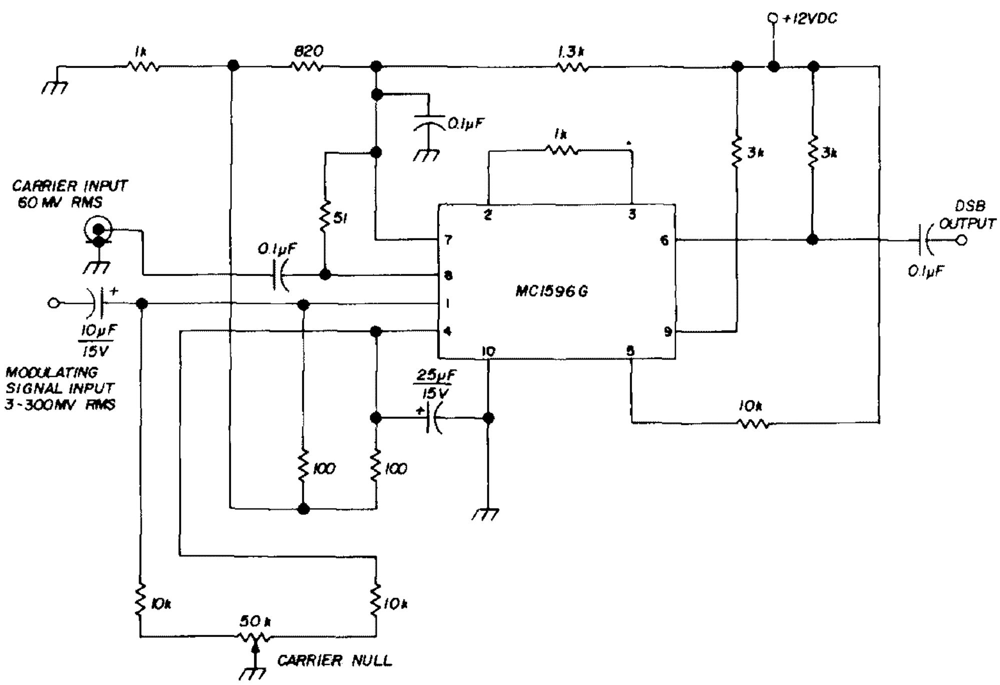



**fig. 3. A balanced modulator using a single 12-Vdc supply.**

To place the modulator in operation, apply dc power and carrier signal. The recommended carrier input level is 60 mV (0.06 V) rms, but this isn't critical and variations in this figure of as much as 50% won't seriously degrade circuit performance. The carrier-null adjustment is set for minimum carrier output.

The audio modulating signal is then applied. This signal should have a value of 300 mV rms on peaks; but again, this level is not critical.

If you have no equipment available to adjust the carrier injection level to 60 mV, apply both carrier and a single audio-tone modulating signal and observe the double-sideband output level on a

voltmeter or scope. As the carrier injection is increased from a very low level (1–10 mV rms) the output will increase as carrier level is increased. Finally a point will be reached where further increase in carrier level causes no change in the output. The carrier level should be set at the point where the output signal just begins to level off.

If only a single 12-Vdc power supply is available, the circuit shown in **fig. 3** may be used. Signal levels and operating instructions are the same as for the circuit in **fig. 2**.

## amplitude modulator

The circuit shown in **fig. 2** can be used for an a-m modulator if the 10k resistors are changed to 750 ohms. This modification gives the carrier-null adjustment the greater range necessary for carrier insertion.

could be used in an ssb/a-m transmitter or in an rf signal generator.

## doubly balanced mixer

A doubly balanced mixer delivers only sum and difference frequency outputs and suppresses both the local oscillator and rf-input frequencies. The MC1596G may be used in this application also.

**Fig. 4** shows a doubly balanced mixer with broadband inputs and a tuned output at 9 MHz. This means that the rf and local oscillator inputs may be any two frequencies with a sum or difference of 9 MHz. The circuit will operate with input frequencies from 160 meters up to 300 MHz. With a 10-meter input signal, the mixer has a conversion gain of 13 dB and a sensitivity of 7.5 $\mu$V for a 10-dB signal-plus-noise-to-noise ratio at the i-f output. At 220 MHz, it has 9-dB conversion gain and a 14-$\mu$V sensitivity.

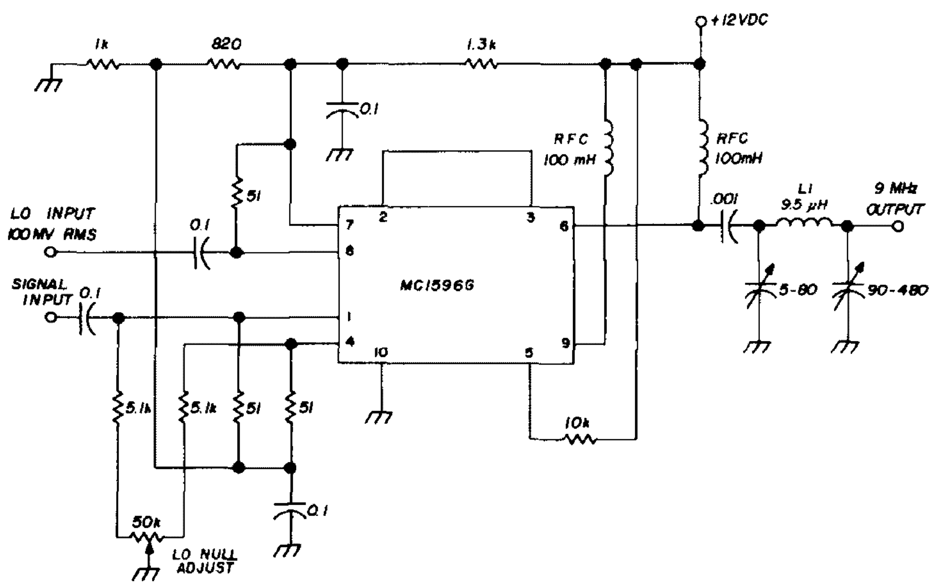



**fig. 4. Doubly balanced mixer with broadband input and tuned 9-MHz output. Circuit operates up to 300 MHz. L1 is 44 turns no. 28 enameled on a Micrometals 44-6 toroidal core.**

The a-m modulator is operated with the same signal levels as the suppressed-carrier modulators described above. It can be used in a system where both suppressed carrier and a-m are required. Simply adjust the 50k pot for either a carrier null or carrier insertion, depending on which operating mode is desired. The modulator

Several variations of this mixer may be used. There are three signal ports, two inputs, and one output. These three ports may be used with any combination of either tuned tanks or broadband coupling circuits. Thus, with broadband circuits on all three ports, the mixer will deliver sum and difference frequency

outputs from audio through vhf. With tuned input and output circuits, much higher conversion gains of 20 to 30 dB may be achieved due to more efficient impedance matching.

the MC1596G, causes the product detector to have only the desired demodulated audio output.

**Fig. 5** shows a product detector circuit. With a 9-MHz ssb i-f input, the

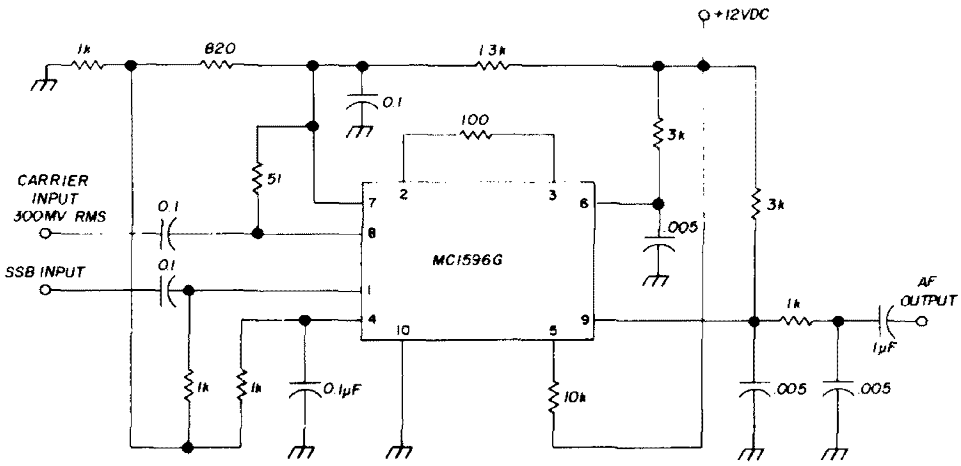



**fig. 5. Product detector with 90-dB dynamic range. The high sensitivity of this circuit lends itself to direct conversion techniques.**

## product detector

An extremely sensitive product detector can be built with the MC1596G.

A product detector is really a mixer with its output in the audio frequency range. The ssb signal forms the rf input,

detector has a sensitivity of 3 μV for 10-dB s + n/n at the audio output. For 20 dB s + n/n, the sensitivity is 9 μV. This means that for an ssb receiver with a 50-ohm antenna input impedance, a 0.5-μV rf input signal would require only 12 dB over all signal power gain from

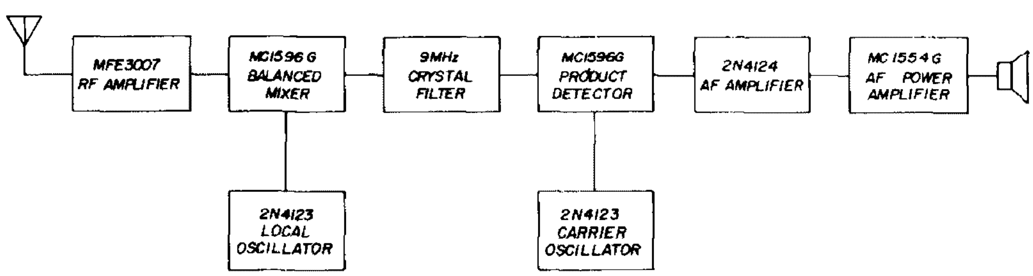



**fig. 6. Block diagram of a receiver with no i-f gain. Sensitivity on 15 meters is less than 0.1 μV for 10-dB s + n/n ratio.**

and the carrier injection signal is the local-oscillator input. The audio output signal is at the difference (audio) frequency between the ssb and carrier frequencies. A low-pass filter that cuts off above 3 kHz is used at the output. This, together with the inherent suppression of both input-signal frequencies provided by

antenna to detector input to produce a demodulated audio signal with 20-dB s + n/n at the detector output.

Of course there would be many other practical limitations on such a receiver such as agc range, audio-amplifier sensitivity, etc. But the point is that detector sensitivity would certainly not be a prob-

lem in any receiver employing the MC1596G as a product detector.

This high sensitivity product detector permits some interesting receiver techniques. For example, the circuit lends itself well to direct conversion.[3,4] Furthermore, it's possible to build a sensitive superheterodyne receiver with no i-f gain. The latter principle has been realized in an hf ssb receiver (block diagram shown in **fig. 6**). The sensitivity of this receiver on 15 meters is less than 0.1 $\mu$V for a 10-dB s + n/n ratio.

The product detector has a dynamic range of 90 dB. This means that it will detect i-f input signals from 3 to 100 mV without significant distortion.

The input-signal-handling capability may be increased at the cost of some decrease in detector sensitivity and gain by increasing the resistor between pins 2 and 3 to 500 or 1000 ohms.

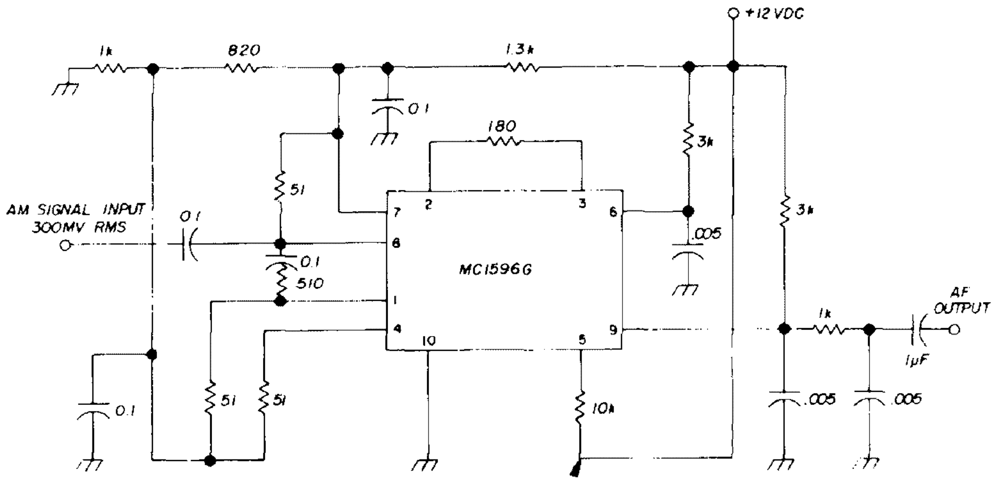



**fig. 7. Optimized a-m detector based on the circuit of fig. 5.**

## a-m detector

The product detector shown in **fig. 5** may also be used as an a-m detector. Thus it can be used as the only detector in an a-m ssb cw receiver.

For a-m operation, simply inject carrier and modulated signal inputs, as for

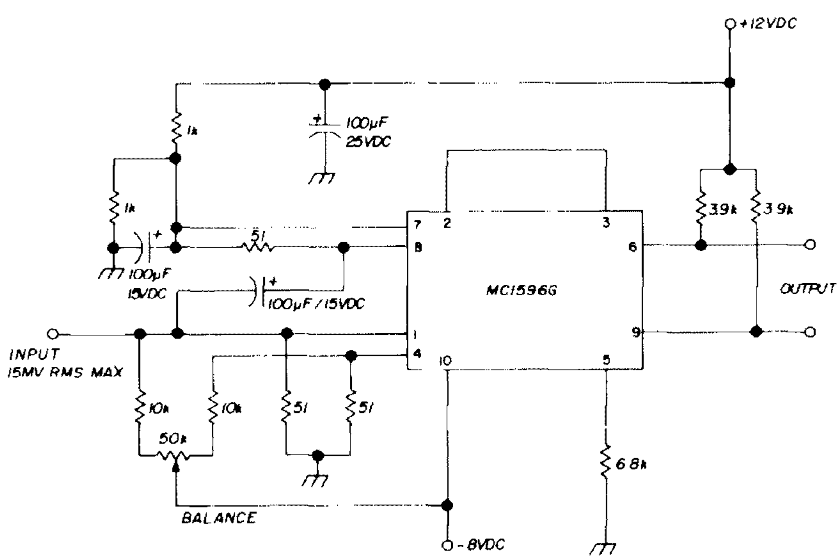



**fig. 8. A low-frequency doubler. Circuit will deliver doubled output from input signals up to 1 MHz; all other frequencies are 30 dB or more below the desired output.**

ssb. The carrier injection is most conveniently obtained from the i-f signal, thus avoiding any drift problems that may be encountered if the carrier is generated locally, as for ssb reception.

range through vhf.

**Fig. 8** shows a low-frequency doubler. The two input terminals are simply ac-coupled, and untuned RC coupling is used at input and output. At input frequencies

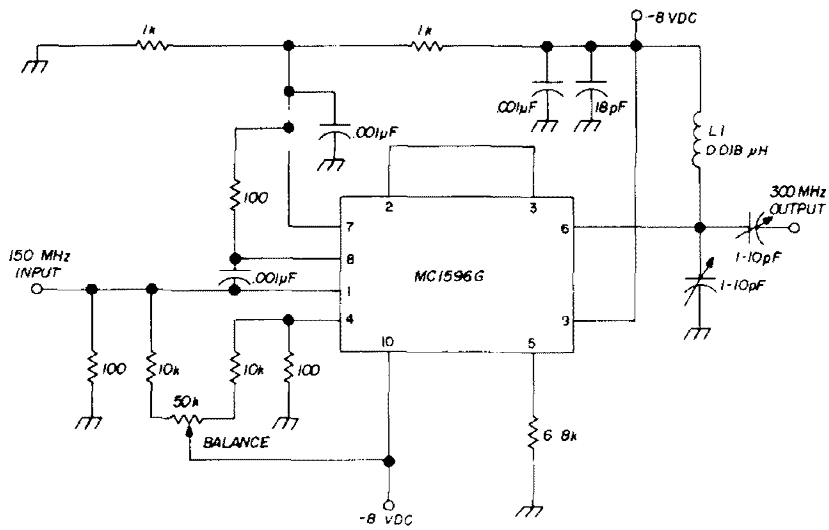



**fig. 9. A frequency doubler for the vhf range. Inductance L1 is 1 turn of no. 18 wire, 7/32-inch I.D.**

Normally, a constant-amplitude local-oscillator (carrier) signal is injected at the local-oscillator input. To achieve this with a carrier signal obtained from the receiver i-f signal, a limiter would have to be used to remove the modulation. However, if a carrier injection level of 300 mV rms is used, the fully modulated signal may be injected directly.

While the product detector shown in **fig. 5** may be used directly as an a-m detector by injecting a 300-mV signal into the carrier input and up to 30-mV signal into the signal input, a few modifications can be added to optimize the detector for a-m. The resulting circuit is shown in **fig. 7**.

## frequency doubler

Injection of the same signal frequency at both inputs produces an interesting result. The sum frequency output is twice the input frequency, and the difference frequency is *zero. Therefore, the output* consists of a single-frequency signal at double the input frequency, and we have a frequency doubler. The MC1596G operates as a frequency doubler without any tuned circuits from the audio frequency

up to 1 MHz, this circuit will deliver a clean doubled output with all other frequencies 30 dB or more below the desired output.

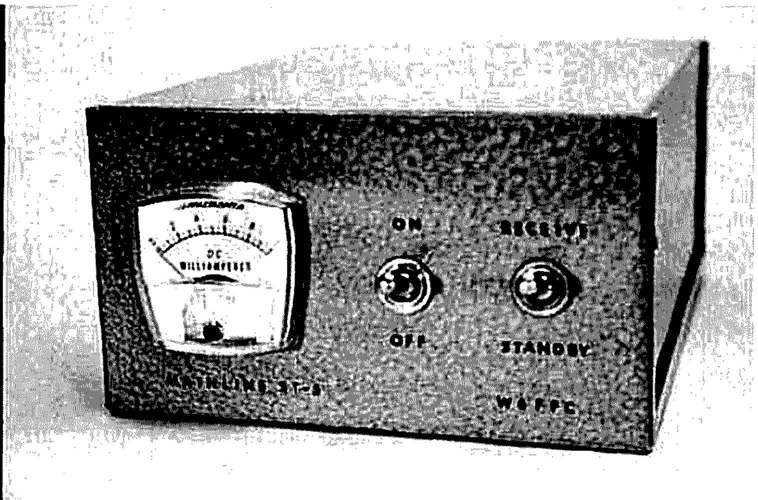
When modified with suitable coupling and bypass capacitors, the basic MC1596G doubler has been used at input frequencies as high as 200 MHz.

Suppression of input frequency and other spurious signals is not as good in the hf and vhf range. Therefore, it may be desirable to use a tank circuit at the output to obtain a cleaner output signal. **Fig. 9** shows a vhf doubler with such a tank. This circuit doubles from 150 to 300 MHz, with all spurious outputs 20 dB or more below the desired output signal.

### references


1. E. Renschler, "Theory and Application of a Linear Four-Quadrant Monolithic Multiplier," *EEE Magazine*, Vol. 17, No. 5, May, 1969.
2. "Analysis and Basic Operation of the MC1595," Motorola Semiconductor Products, Inc., Application Note AN-489.
3. Hayward and Bingham, "Direct Conversion—A Neglected Technique," *QST*, November, 1968, p. 15.
4. Richard S. Taylor, "A Direct-Conversion S.S.B. Receiver," *QST*, September, 1969, p. 11.


**ham radio**




# the Mainline ST—5 rtty demodulator

This basic building block demodulator featuring linear IC's can be used for future expansion of your rtty station

Irvin M. Hoff, W6FFC, 12130 Foothill Lane, Los Altos Hills, California 94022

Many newcomers to rtty have complained that a current yet simple demodulator hasn't been published for them to build. The W2PAT unit in the ARRL handbook is nearly 15 years old. In 1964 an attempt was made to replace the W2PAT design with a modestly priced updated unit, the TT/L.[1] This design, together with the subsequent TT/L-2,[2] is now the standard of the serious rtty enthusiast. However, the original goal was missed by a country mile, since the TT/L-2 costs over $160 just for parts and has 14 tubes.

The ST-3[3] was a successful solid-state design that introduced integrated linear operational amplifiers to rtty. It was still moderately complex, however, and fell short of the goal to supply the beginner with something that could be built in a few hours.

## the ST-5 demodulator

While developing a unit based primarily on ICs to replace the TT/L-2, a very simple modulator with great potential was developed: the ST-5. As with any simple circuit, the cost of the power supply is out of proportion with the rest of the unit. At current prices, the ST-5

costs only $14.50 less loop supply ($8) and a plus-minus 12-volt supply ($11).

The total cost of $33 is not overly impressive until you realize this unit can, if desired, be used as a building block for the more exotic ST-6, which will be published later in the year. Almost every component used here can be used in that unit. The ST-5 is a basis from which the beginner can expand—it's not just a collection of parts that will find no further use when he is ready to broaden his horizons to more sophisticated equipment.

## features

The ST-5 uses two operational amplifiers (fig. 1). One is an audio limiter, and the other is a trigger stage to drive the keyer. It has a 175-volt loop supply of the same type used in the TT/L, which provides plus-minus voltages for keying a transmitter and also features narrow-shift cw identification. Finally, the ST-5 has a symmetrical plus-minus 12-volt power supply.

**Rear panel of ST-5 shows jacks for audio, fsk and loop power. Enclosure is 6 x 8 x 3½ inches.**

## limiter

The 709C op amp has over 90-dB gain and is good to over 10 MHz. It makes an ideal limiter. The zener diodes on the input don't assist in the limiting; they merely protect the 709C against damage in the event of excessive audio input (hardly likely but worth the protection). The limiter puts out square waves and is so powerful it starts working on input signals as low as 200 $\mu$V. The 25k pot merely balances the small offset input voltage for maximum gain. This voltage varies slightly from one unit to another, so a control pot was added rather than a fixed resistor, which many units use.

## discriminator/detector

It's difficult to use the same value inductor with different capacitors and expect to obtain two similar filters of equal characteristics. To get similar bandwidth, voltage output, noise response, etc., some loading is necessary. Most simple demodulators merely balance the voltage or ignore all the problems completely. Without belaboring the point, it's not a simple job to get all these factors to balance suitably; but it *is* possible, and the Mainline units all have filters that have been designed with care.

The ST-5 offers a choice of the 2125-2975 mark and space tones (considered standard), or the 1275-2125 low tones necessary in some modern receivers. (Actually nearly all these receivers respond beautifully to 2975 tones and higher, but a new bfo crystal is needed.) The best results come from the 2125-2975 tones, since the two frequencies are only about 28% apart while the 1275-2125 tones are 40% apart; thus it's a more difficult job to separate the harmonics and achieve proper filter design.

The detector features full-wave rectification for most efficient filtering of the dc ripple remaining after the audio has been rectified. A simple RC low-pass filter removes the remaining audio component.

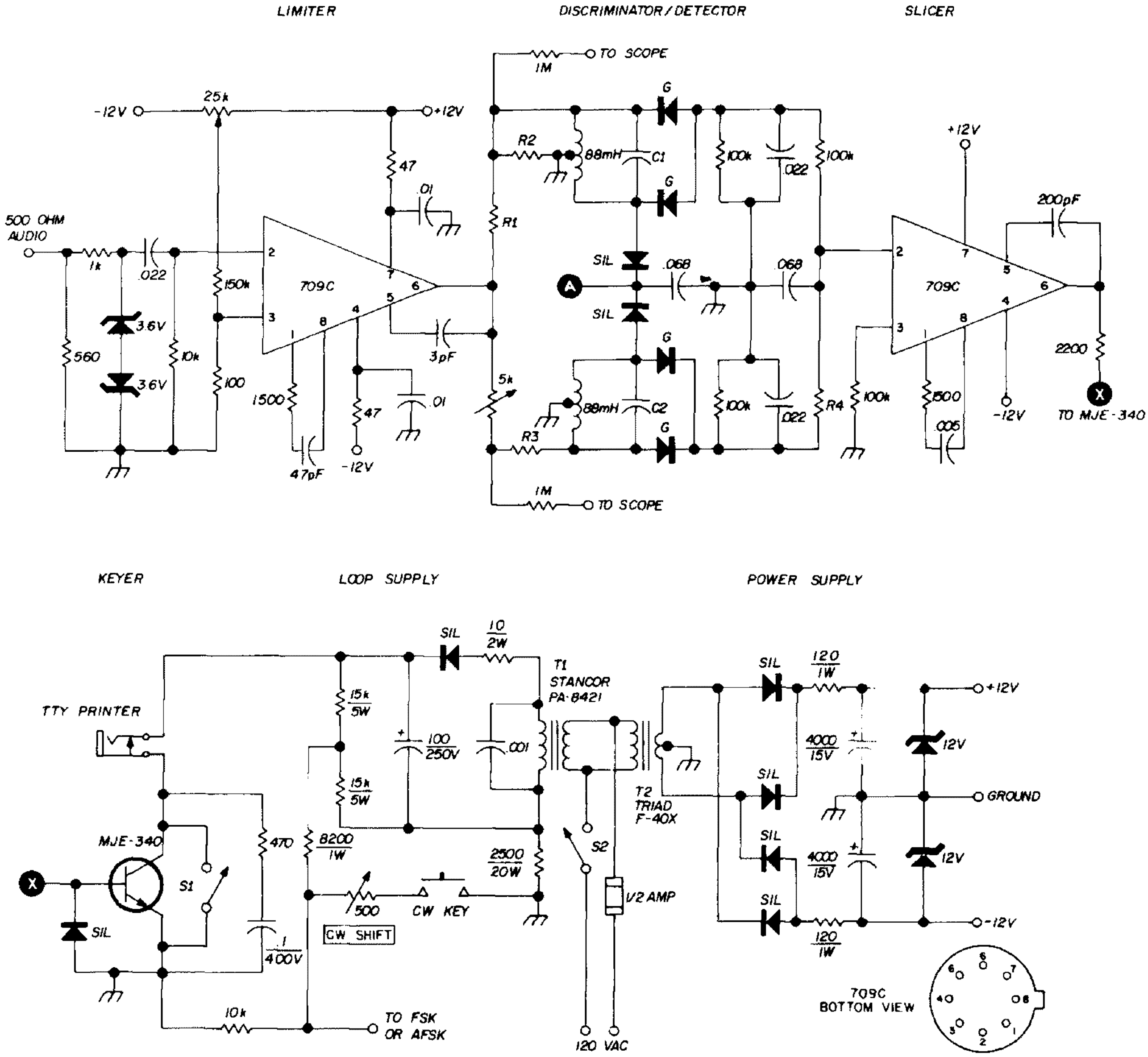



**fig. 1. Schematic diagram of the Mainline ST-5 RTTY demodulator.**

## slicer

The slicer takes the small voltages from the filters and changes them to roughly +10 volts for mark and -10 volts for space, regardless of the original amplitude. This in reality is a dc limiter, as a signal as small as a 100 μV or so will cause the unit to saturate completely, either plus or minus, depending upon the polarity of the applied signal voltage. The unit has so much gain that at the crossover point, a change at the audio input as small as one or two Hz will cause this trigger stage to flip from +10 to -10 volts. This is another way of saying shifts as low as 3–4 Hz could be copied on the ST-5 if tuned in properly.

## keyer stage

A 300-volt Motorola 25W transistor selling for $1.06 is used. The normal loop-supply current for tty machines is 60 mA. This transistor has a large amplification factor and acts like an on-off switch. When on, the power consumed in the transistor is only 0.012 W; so in the ST-5 there's no way you could ever damage that transistor.

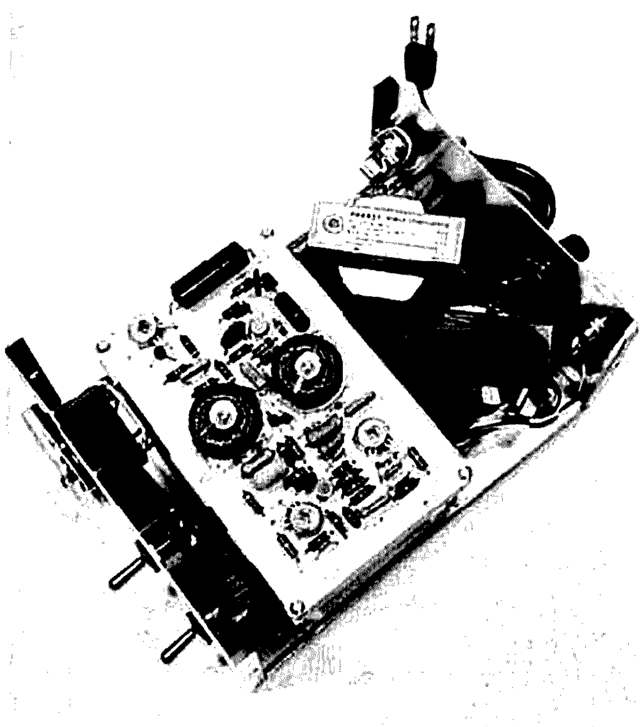
An RC network in the MJE-340 collector takes care of the back emf developed by the inductance of the selector magnets in the printer during the transition from space to mark. The transistor is biased off during space. A diode in the base circuit keeps this negative voltage below the

point at which the base-emitter junction would be reverse-biased.

## loop supply

Here comes $8 of the $33 total right now. This unit uses the well-known "floating loop" I developed for the TT/L. As you go from mark to space, the voltage at the fsk output switches from negative to positive. This offers excellent keying characteristics for the transmitter, and provides a simple method of keying ssb transmitters needing conduct on mark instead of conduct on space, such as the Collins S-line. The Hallicrafters (and others using a 9-MHz heterodyne scheme with a vfo running from 5.0 to 5.5 MHz) needs both systems. If you are "upside down," merely reverse the diode in the fsk system. Few (if any) other systems offer this potential. The narrow-shift cw identification system can be set appropriately with the 500-ohm pot. If you are using a transmitter that conducts on mark and can't get suitable cw shift, try putting the connection to the 500-ohm pot on the other end of the 8.2 kilohm resistor. One of these two places has always been adequate in the past.

The 2500-ohm resistor sets the loop current, which is in no way critical. Unless more than 10 mA in error from 60 mA (you may have used a different transformer and need a different resistor), don't bother changing anything.



**Printed-circuit boards hold all the components except the two transformers and the control switches.**

## loop transformer

The Stancor PA-8421 is far from cheap. However, I've never found a more suitable transformer at any lower price. Don't be alarmed at the 50-mA rating, which requires some explanation. The primary is capable of handling about 20 VA in the secondary. Since there's also a 6.3-volt winding rated at 2 A, this is almost 13 of that 20 VA. That only leaves about 7 VA for the high-voltage winding, or roughly 50 mA. However, if the filament winding is not used (and I don't use it), then the entire 20 VA is available to the high-voltage winding. This represents around 160 mA. So don't be alarmed at the 50-mA rating. You could pull twice that in this circuit and it wouldn't tax the transformer. Don't worry if the transformer gets warm; all transformers get warm. It's when you burn your hand on them that you have to watch out. I've had a loop supply similar to this running in the TT/L twenty-four hours a day for six years, and others throughout the country are doing the same.

## power supply

The 709C op amps will take up to $\pm 18$ volts. If you wind up with more than the indicated $\pm 12$ volts, but less than $\pm 18$ volts, think nothing about it. You can lower the voltage by increasing the value of the resistors if desired. The plus voltage goes up or down at the same rate as the minus voltage, since both voltages are supplied by the same transformer. The op amps use symmetrical voltages. The 4000-$\mu$F capacitors are Sprague type 39D at $2.43 each. Other large-value brands may be used, and I suggest you use at least 3,000 $\mu$F for this purpose if substituting.

## standby switch

When S1 is closed, the unit is placed in mark. When S1 is opened, the printer can follow whatever is fed into the limiter from the receiver.

As explained previously, the unit has so much gain that a signal as small as 3–4 Hz can be copied if tuned correctly; this is called straddle tuning. However, for 170-Hz shift you may wish to add a switch that changes the space filter to the new frequency. **Fig. 2** shows the way this would be accomplished if using the normal 2125-2975 tones, and fig. 3 shows the circuit for the low tones of 1275-2125. This is merely an expedient and doesn't result in proper filter balance, but it provides good 170-shift reception with the switch closed, or normal 850-shift reception with it open.

## tuning indicator

Provisions are provided for connections to the vertical and horizontal amplifiers of a scope (fig. 1). It is customary to connect the mark signal to the horizontal amplifier and space signal to the vertical

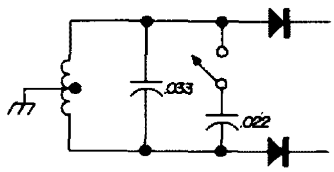



**fig. 2. Switching circuit for adding 170 shift to space filter for 2125-2975 tones.**

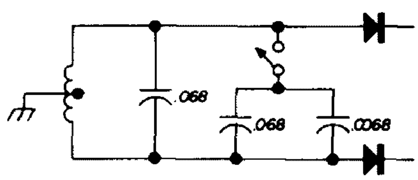



**fig. 3. Switching circuit for adding 170 shift to space filter for 1275-2125 tones.**

amplifier, although many reverse this method.

Most people prefer a scope indication, but an excellent tuning indicator is provided at point A (fig. 1). A voltmeter connected to this point will give equal voltage indication for mark or space. With rtty signals the meter should stand still. If it doesn't, retune the receiver until it does. If straddle tuning a signal, the meter may read less than normal, although it won't move. This is normal and merely indicates the shift being copied is not the correct shift for the filters you're using.

Fig. 4 also shows how a 0–1 mA meter may be added. An inexpensive npn transistor is used, such as the MPS-3394, although any npn transistor would be satisfactory here. The capacitor merely dampens the meter so it doesn't flip around too violently. If your meter is too damped, remove the capacitor or try a smaller value. This was suitable for the inexpensive imported meter used in my unit.

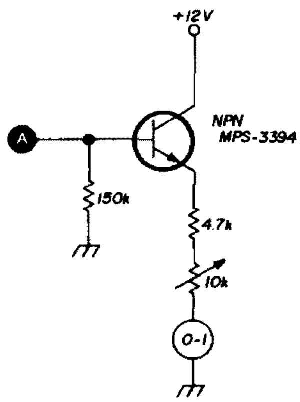



**fig. 4. Simple tuning indicator uses an inexpensive 0-1 mA meter.**

## the transmitter keyer

Fig. 5 shows a typical fsk keyer for installation in the transmitter. The components can be mounted on a small terminal strip and placed near the vfo tube under a convenient mounting screw which also serves as a ground return. The trimmer is connected to the cathode of the vfo tube and the tube replaced in its socket; thus, no changes of any type are made to the transmitter and its resale value is not affected. There should be room for several keyers if you wish to have the convenience of both 170 and 850 shift.

Although a 3-12 pF trimmer is shown in **fig. 5**, some transmitters only require a 1.5-7 pF trimmer. It is suggested that you do not substitute for the 1N270 diode as it is superior to most other types in this application.

If your signal is reported as "upside down," reverse the 1N270 diode. If you do not obtain sufficient cw shift with this connection (conduct on mark), the

500-ohm cw-shift pot should be connected to the opposite side of the 8.2k resistor at the junction of the two 15k resistors (fig. 1).

## components

The 709C op amps are supplied by various manufacturers including Signetics, Fairchild, and Motorola. The prices in the order named are $2.62, $2.65 and $2.80 as of this writing. Prices are constantly being reduced as devices become available from more companies. When I first started working on a super deluxe demodulator in the fall of 1967, I paid over $10 each. Now they're too cheap not to use.

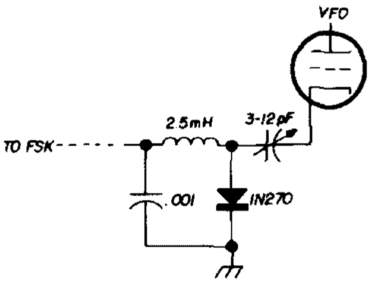



**fig. 5. To add fsk to literally any transmitter, connect a 3-12 pF trimmer to the cathode pin of the vfo tube.**

The Motorola unit can be purchased through most distributors, including Allied and Newark. The Fairchild unit can be mailordered from the firms below.* Include about $1 extra for handling and postage; the surplus will be refunded. Specify the TO-5 can, as this is easier to work with than the dual in-line 14-pin type (same cost).

The diodes marked G in fig. 1 are 1N270 germanium at 32¢ each. Those marked SIL are most any silicon type, such as the 1N2069. The one in the loop supply should, however, be a minimum of 400 volts PIV. Fifty-volt PIV is suitable everywhere else.

*Hamilton Electro Sales, 340 East Middlefield Road, Mountain View, Calif. 94040 and G. S. Marshall Co., 732 North Pastoria Avenue, Sunnyvale, Calif. 94086 (also carries Signetics). If buying Motorola version, ask for the MC-1709CG. Texas Instruments 709 op amps are $1.50 each (or 7 for $10) from HAL Devices, Box 365H, Urbana, Illinois 61801; ask for SN72709L.

If you don't wish to spend the money for the zeners on the limiter input, you can substitute regular silicon types as shown in fig. 6. These start clipping at 0.6 volt, however, and offer an inferior form of limiting, although they more than adequately protect the input to the limiter from excess voltage. The zeners are a much better choice.

The 88-mH toroids are available from various sources for about 40¢ each.† They're wired in series for 88 mH, and the junction of the two windings is grounded.

If you have an accurate means of determining the frequency, you can tune the filters by removing turns of wire from each of the two sections concurrently to keep the turns ratio in the two windings the same. One turn from each of the two windings will increase the frequency about 6 Hz at the 2125 frequency, for example.

Use Mylar capacitors, such as the Sprague Orange Drop. Twenty-five-volt capacitors are adequate, but you'll probably wind up getting 200V types. They are only 15–21¢ each.

The pots can be the inexpensive 39¢ Mallory PC board MTC types. Other power transformers may be used, but the Triad F-40X is an excellent buy.

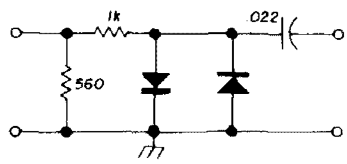



**fig. 6. Ordinary silicon diodes may be substituted for zener diodes at the input, although limiting is somewhat inferior.**

## printed-circuit board

The printed-circuit boards shown in the ST-5 in the photographs hold all of the components except the two transformers and the control switches. This greatly enhances construction, and at the same time makes it possible for nearly anybody

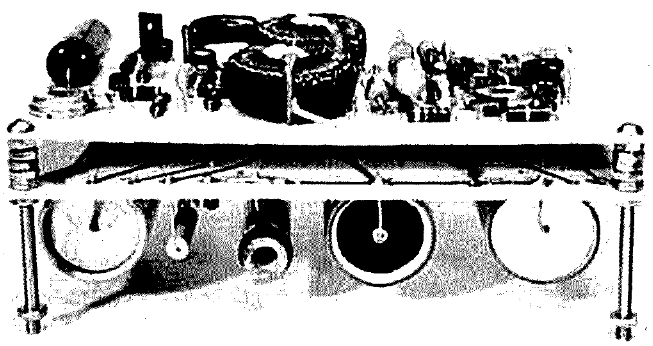
† An excellent source for 88-mH toroids is M. Weinschenker, Post Office Box 353, Irwin, Pennsylvania 15642. He will send you five 88-mH toroids for $1.50, postage paid.

to build an extremely nice-looking unit. The printed-circuit board includes one section for the power supply and another for everything else. The board may be split down the middle and the two sections mounted back-to-back as I did in my unit, or the board may be left intact and used with a more shallow chassis.*

## adjustment

With no input signal, or with the input grounded, adjust the pot on the limiter for zero volts dc at pin 6. If this isn't possible, you'd better write me and explain thoroughly, as you probably ruined the op amp somehow. By the way, unless you get too much voltage on pins 2 or 3, like the full power-supply voltage, or get the plus-minus hooked up backwards, it's very difficult to ruin these things. By following even the most elementary construction practices, you'll have no problems with the 709C.

In this model the printed-circuit board is cut in two sections and mounted back-to-back.



*A printed-circuit board for the ST-5 RTTY terminal unit is available from Stafford Electronics, Inc., 427 South Benbow Road, Greensboro, North Carolina 27401. The undrilled board is $3.00; with critical holes drilled, $3.75; complete, ready to mount components, $6.50. A complete kit of components (less circuit board), including ICs, transistors, diodes, resistors, capacitors and toroids is available for $37.50 from HAL Devices, Box 365H, Urbana, Illinois 61801.

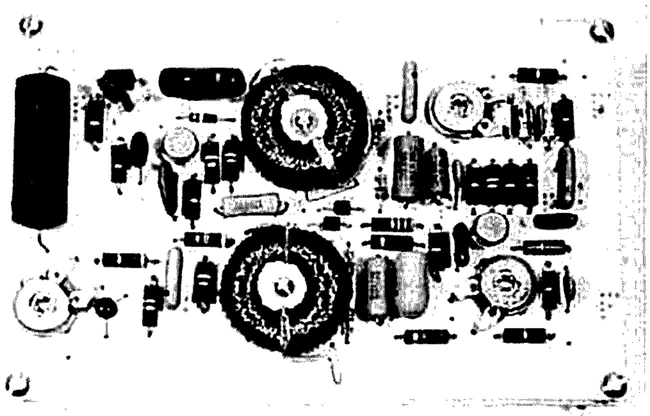
After balancing the limiter for zero volts output, connect the receiver and tune to maximum mark and note the indication on your tuning indicator (fig. 4) or on a voltmeter connected to point A. Tune to space on the receiver and again note the reading. If the indications are not the same, adjust the 5k pot on the limiter output until they are. You have now finished all the adjustments and they should require no further attention at any time unless you switch to 170 shift, for instance. In this event you may or may not want to reset the filter balance pot. I suggest you leave it for the 850 setting and take what you get on the 170-switch position, as this is a somewhat artificial method of getting good 170-shift reception.

When transmitting be certain to first close the standby switch or you can get feedback, which will produce errors similar to those you would get when using a microphone if you didn't turn off the speaker.

## other op amps

The 709C is to other op amps what the Ford V-8 was to other automobiles. It not only led the way; it's still in use. The 709C was (and is) one of the cheapest ICs of its type available. One would gain very little and stand to lose a lot by trying to substitute other units. The 741

Component layout on the main section of the printed-circuit board.



and 748, for example, have a bit more gain, higher input voltages, and require no frequency compensation. They cost $4.85 each, but their biggest disadvantage here is that they're not at all suited as audio amplifiers. At 2 kHz they have only 30-40 dB gain and make a poor audio limiter compared with the 709C. So unless you know what you're doing, stick to the 709C.

## the Mainline ST-6

This demod will be published as soon as possible. The ST-6 uses the same limiter and slicer as the ST-5 and the same keyer and loop supply. It uses the same power supply to which has been added some regulation. It has a total of 7 op amps and 9 transistors. Also featured are an antispace circuit, an active 3-pole Butterworth low-pass filter, autostart with delayed motor control, and optional features such as bandpass input filters for

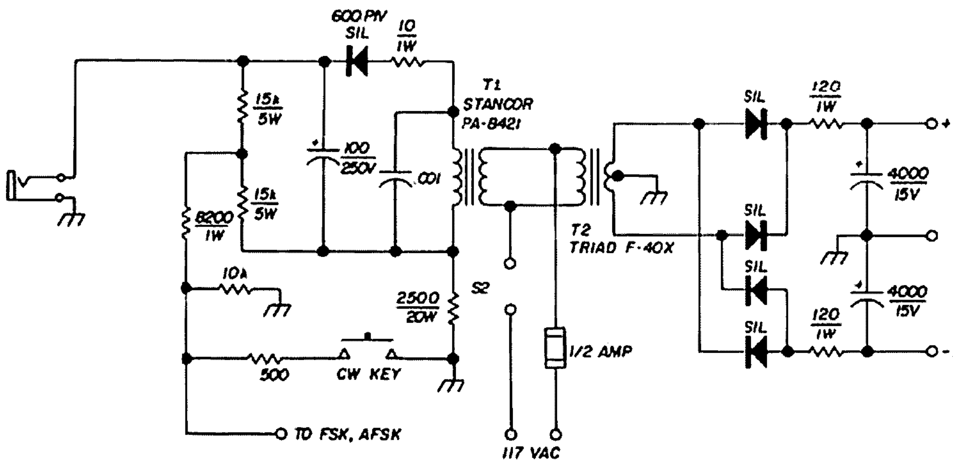
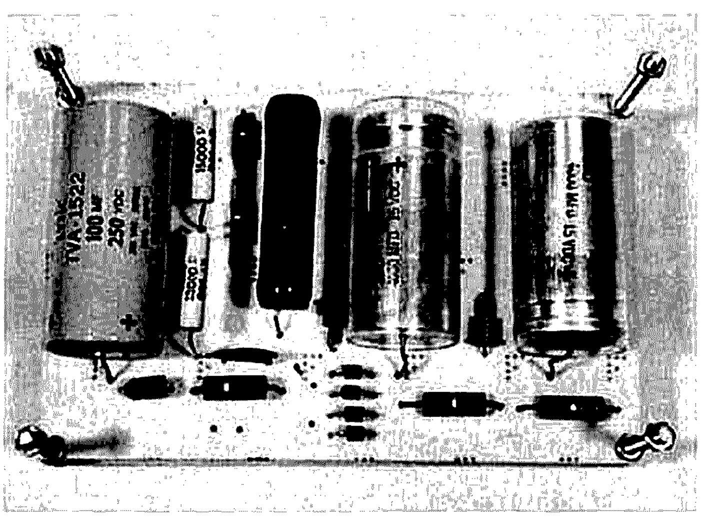



fig. 7. Although this unregulated power supply may be used with the ST-5 RTTY demodulator, the circuit shown in fig. 1 is recommended.

layed motor control, and optional features such as bandpass input filters for either 170 or 850 shift, fast-slow autostart, etc. This is only mentioned at this time to illustrate that if one builds the ST-5, the same parts may be used later for the more exotic ST-6 if you wish to expand your station.

Component layout on the power-supply section of the printed-circuit board.



## conclusion

The ST-5 was designed as a simple but highly effective rtty demodulator using the best of currently available concepts. It should be a very popular unit for some years to come, as it's impossible to imagine at this time how any additional performance could be made available—it's already ridiculous to talk in terms of 90+ dB amplification. Only a completely different concept of rtty processing could outdate the ST-5, and that seems quite unlikely to occur until we all get computer terminals in the shack.

### references


1. *RTTY,* November, 1964; also *QST,* August, 1965.
2. *RTTY Journal,* September, 1967; also *QST,* May, June, 1969.
3. *RTTY Journal,* September, 1968; also *QST,* April, 1970.


ham radio

# an fm receiver

## for two meters

Conservative design, solid-state construction, readily available parts— all add up to a really solid vhf fm receiver

Ron Vaceluke, W9SEK and Joe Price, WA9CGZ

**Within the last several years,** hams have been giving increased attention to vhf fm operation. Although this mode has been used by hams for over 15 years, it is just beginning to be widely accepted. One of the reasons is because of the release from commercial service of the older, wide-deviation, tube-type gear. Commercial users had to go to a narrower deviation to open up more frequencies. One of the authors was on wideband fm 16 years ago using what was at that time fairly current equipment. After all these years, the same type of gear is still widely used. With today's techniques and technology, much of the older tube-type equipment is obsolete. The tragic part is that since the demand for vhf fm equipment has increased, the law of supply and demand has raised the prices of these antiques out of proportion to their usefulness.

Lately several distributors have been selling imported fm gear designed for ham use. Although we haven't tried this equipment, it probably works quite well. However, we felt that it should not be necessary to get a second mortgage on our homes to be able to purchase hobby equipment. A problem that can exist with imported electronic gear is the availability of adequate repair parts when needed. With all of these objectionable possibilities in mind, it was felt that the best approach to vhf fm operation was of the do-it-yourself type. The solid-state, conversion receiver described here is the result.

In establishing our design criteria we felt that it would be wise to spend a few extra dollars, build more than just a basic receiver, and do the job right the first time to obtain good rather than mediocre operation. In looking over readily available semiconductors, best use was made of both discrete and integrated devices.

The first version of the receiver was built on punched board, using eyelets at every point of component entry. Although this technique is adequate for initial design, it leaves something to be desired as far as a finished product is concerned. The second version was made using printed circuits.* All parts are readily available and standard. The i-f transformers are imported, but are a standard type used in portable radios. These are also available from J. W. Miller. Don't overlook those defunct transistor radios at hamfests, etc., as a source of transformers. The coil forms in the rest of the circuit are made by Cambion (avail-

*A set of G-10 epoxy PC boards, tinned, drilled, and with swaged terminals is available for $9.50 P.P. A PC board for the power supply, G-10 epoxy, drilled and tinned is available for $2.50 additional. Order from RMV Electronics, P. O. Box 283, Wood Dale, Illinois 60191.

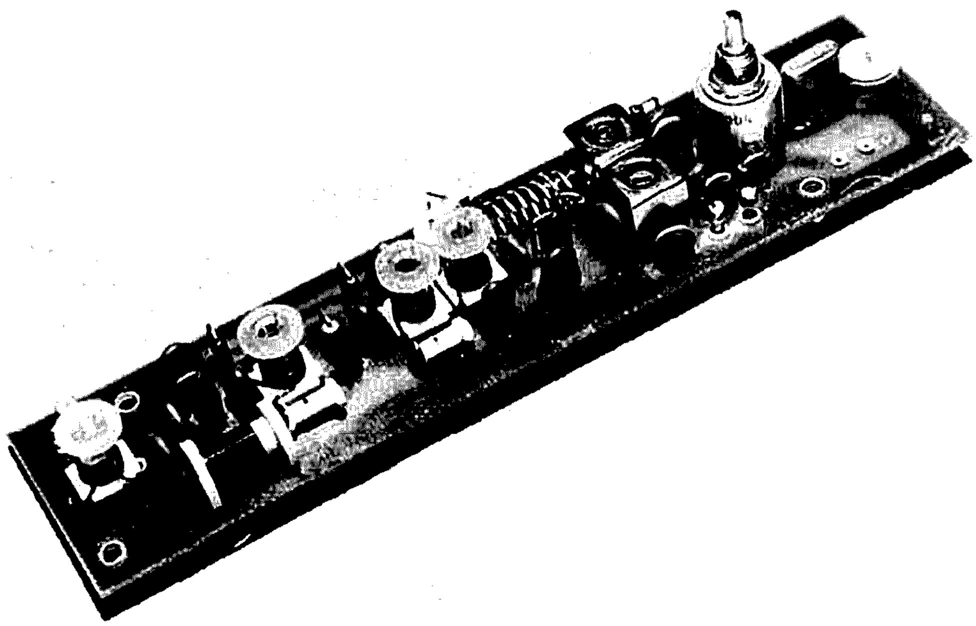
able at Newark Electronics, etc.) and can be shielded by the same size coil shields used on the i-f transformers.

## the front end

For good cross-modulation and overload characteristics, fet transistors are used in the front end (fig. 1). The first rf amplifier, Q201, is a conventional, grounded-source, neutralized MPF-102. Other types may be somewhat better; the MPF-105 or MPF-107, for example.

variety is recommended since the larger ones have too much inductance. In the printed circuit version we used an ALCO MRA-3-3S switch. The pins went right through the board and to the foil.

The first mixer is a MPF-102 fet. The signal and local oscillator frequencies are fed to the gate. The source resistor was chosen empirically for the proper amount of local-oscillator injection. The mixer output is 10.7 MHz and is taken off by a link on T201.



**View of the PC front end. At the left is the rf input and Q201, followed by Q202. At right center is Q203 mixer, and at the right end of board is the oscillator. The board will hold up to three crystals although only one is shown here.**

A grounded gate stage, Q202, is used as the second rf amplifier to simplify construction and for ease of adjustment (no neutralization).

The first local oscillator consists of a crystal oscillator, Q204, operating at 45+ MHz, which is tripled to approximately 135 MHz by a 1N914 diode. The crystal frequency can be determined as follows:

$$\text{xtal freq} = \frac{\text{operating freq} - 10.7\ \text{MHz.}}{3}$$

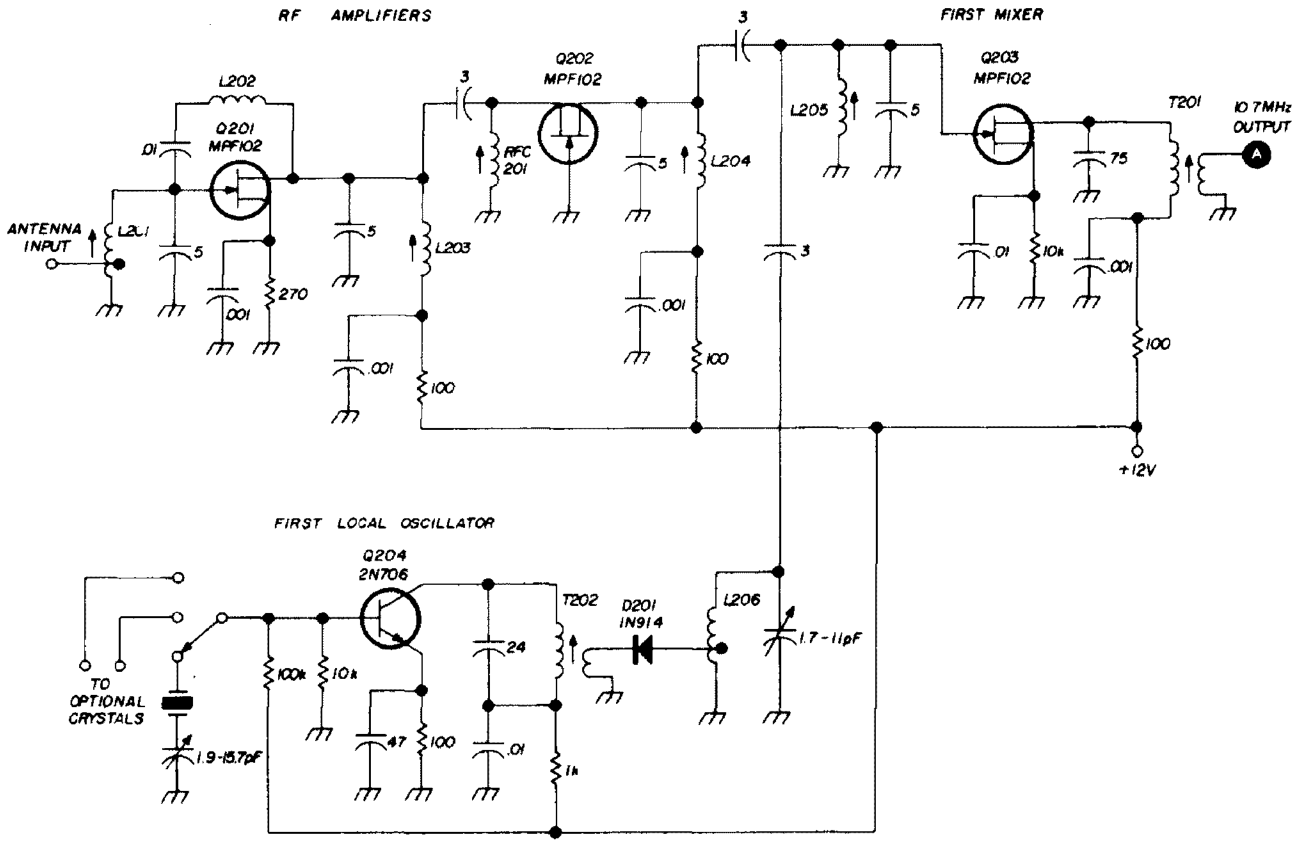
A trimmer capacitor in series with the crystal allows frequency zero adjustment. Be sure to use short leads from the base of Q204 to the crystal, or it won't oscillate. If a switch is used, the miniature

## main i-f board

The output of the first mixer is fed to the gate of Q101, the first i-f amplifier (fig. 2). Link coupling from the first mixer provides an impedance mismatch, which prevents Q101 from oscillating without neutralization.

Due to the low Q of the transformers in the mixer and first i-f amplifier outputs, there may be a problem with an image 910 kHz from the desired signal. If you're in an area with lots of activity, or discover later that this problem exists, an external filter can be added between the converter and first i-f.

The output of Q101 is transformer-




| | |
|---|---|
| L201 | 4½ turns no. 18 on 0.210" diameter slug-tuned coil form (CTC 3624-4), tapped at 1¼ turns |
| L202 | 18 turns no. 26 on 0.214" diameter slug-tuned form (CTC 3104-4) |
| L203, L204 | 4 turns no. 18 on 0.210" diameter slug-tuned form (CTC 3624-4) |
| L205 | 3 turns no. 18 on 0.210" slug-tuned form (CTC 3624-4) |
| L206 | 7 turns no. 20, 3/16" diameter, air wound, tapped at ½ turns |
| RFC201 | 8.2 µH (Nytronics DD-8.20 or equivalent) |
| T201 | CTC 3624-2 coil form wound full with no. 32, 8-turn link, shielded (winding area of CTC 3624 is 0.183" long, 0.210" diameter) |
| T202 | CTC 3624-3 coil form wound full with no. 22, 3-turn link, shielded (see coil form info under T201) |

**fig. 1. Front end of the vhf fm receiver. First rf stage neutralization adjustment (L202) will depend on individual transistor characteristics. Neutralization isn't required in grounded-gate second stage.**

coupled to the base of the second mixer, Q102, where it is mixed with 10.245 MHz from the second local oscillator, Q103. The difference frequency, 455 kHz, is fed to a three-stage filter consisting of T102, T103 and T104. The selectivity depends on the coupling capacitors and also the tuning of the transformers.

Other types of filters are available that have better bandpass characteristics, but their cost and limited availability prohibit their use. By using i-f transformers as a filter, they can be adjusted for either wide- or narrow-band use. The i-f transformers are the type used in transistor radios.

The second i-f amplifier, U101, is an RCA CA3012 integrated circuit, which is low priced and has approximately 65 dB gain with good limiting characteristics. Because of the very high gain, the lead dress is important, and the bypass capaci-

tors should have short leads to prevent self-oscillation. A problem developed on the original printed circuit layout with this stage due to ground loops, which caused U101 to oscillate.

Transformer T105 couples the i-f amplifier output to the fm detector U102, which is a Sprague ULN 2111A. This integrated circuit contains an i-f amplifier, limiter, fm detector, and an audio stage. Other IC fm detectors are on the market, but this unit doesn't require an expensive transformer and provides excellent a-m rejection.

The audio output of the fm detector feeds the audio gain and squelch pots. Transistor Q105, a noise amplifier, drives Q106, the noise detector. Transistor Q107 is a dc amplifier, which is driven by the noise detector and is used to turn the audio squelch gate Q104 on and off by biasing the emitter-to-base junction on and off.

Audio from the squelch gate drives the audio output stage, which is a relatively new integrated circuit from Motorola. The MFC4000 is inexpensive, small, and provides 250 mW output. While this is not quite enough for mobile use, it provides more audio than you can stand with a 4- or 5-inch speaker in normal room conditions.

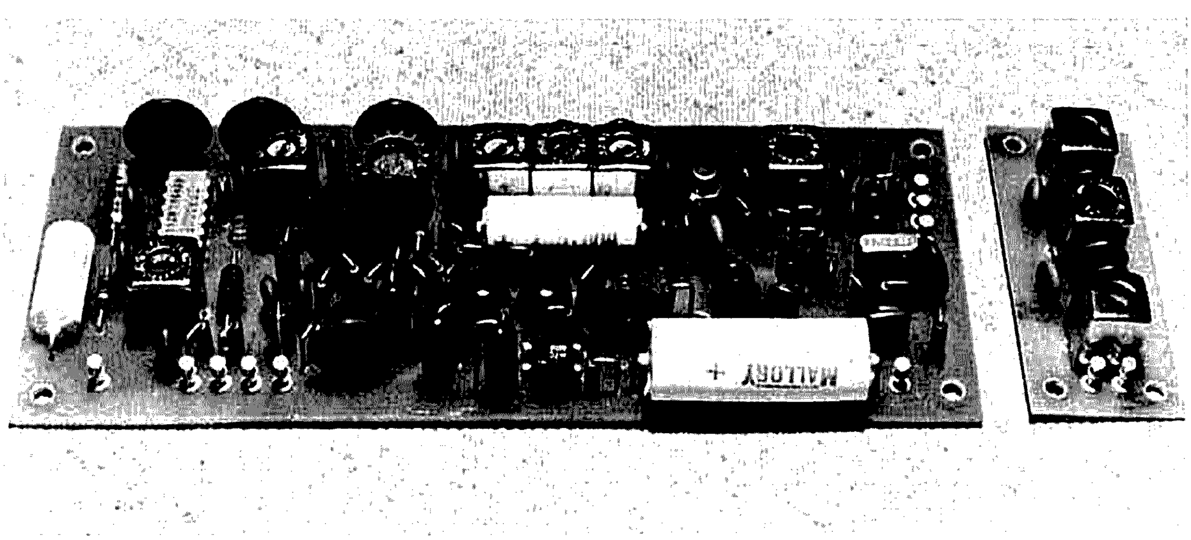
## power supply

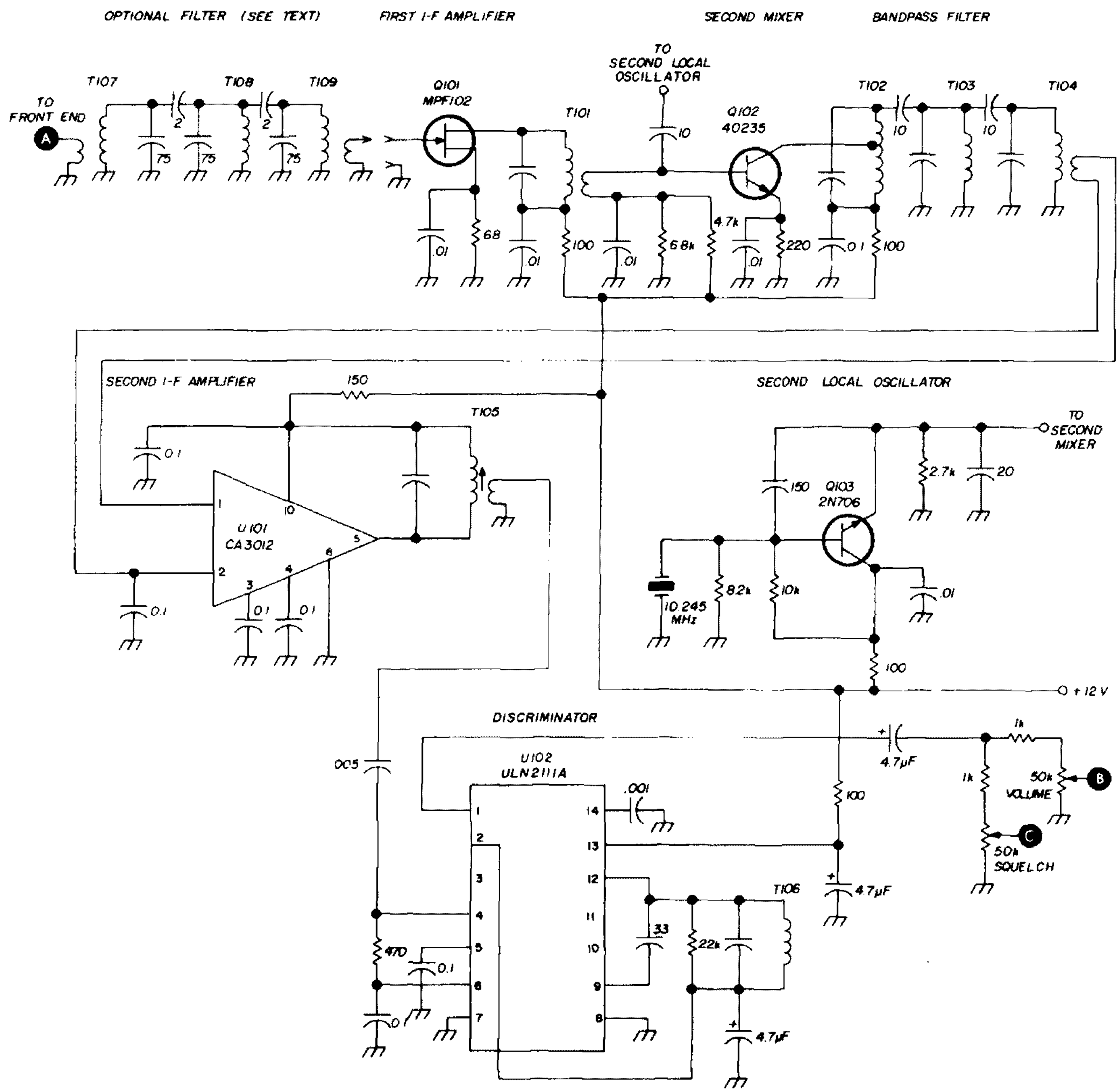
The entire receiver operates from a 12-volt source. Audio output stage U103 requires 9 volts, which is obtained from a zener. This method was used rather than a dropping resistor, because the current to U103 varies widely and so would the voltage drop. If the receiver is used for mobile work, some additional filtering of the battery supply may be desirable. The amount can vary from car to car. To use the receiver in the home, a small ac power supply can be made. One of the authors has poor line regulation, so an electronically regulated supply using a Motorola MC1460R integrated-circuit regulator was built to power the receiver. The receiver draws approximately 60 mA when squelched (no audio) to around 150 mA on speech peaks with the audio gain wide open. As far as mobile is concerned, since the power requirements are so low, many hours of listening can be had without straining the car battery.

## alignment

Although an expensive fm signal generator of the type used for commercial radio work would be nice, it's not at all necessary. The i-f board can be initially tuned using any 10.7-MHz signal generator with a-m modulation. Some may question this; however, if the generator output is kept low as well as the percentage of modulation, the receiver will respond if the level is low enough to prevent limiting.

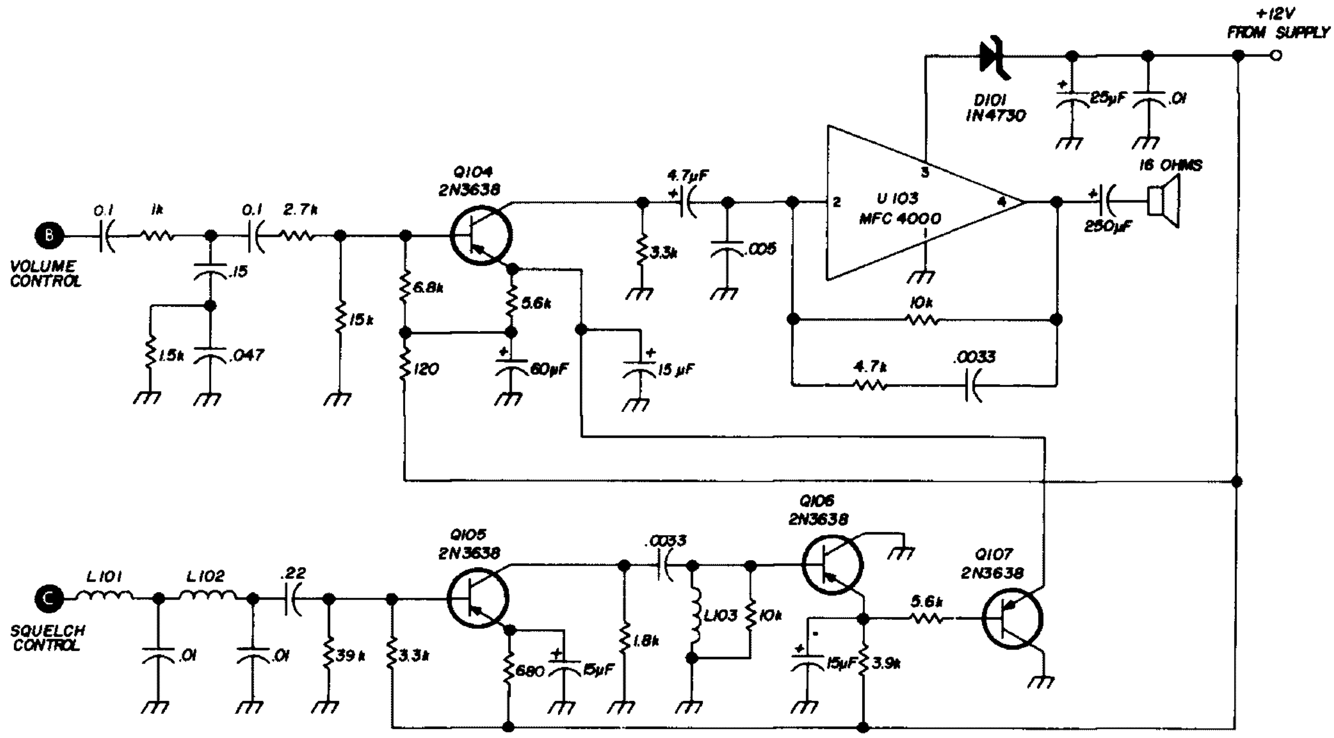
**The large board (approximately 6 x 2-7/16 inches) holds the i-f amplifiers, discriminator, audio and squelch circuitry. The smaller board at right holds T107, 108 and 109. The layout is compact but not miniaturized.**






To start, inject 10.7-MHz to the gate input of transistor Q101. Since we need a dc return, (normally provided by the link on T201) a 2.5-mH rfc should also be connected from the input to ground. The generator coupling should be through a capacitor of approximately 100 pF. Peak all transformers for maximum audio output, but be sure to keep the generator output low to prevent limiting. Next, connect the front end to the i-f board. (Be sure to remove the rfc.) Place an rf probe connected to a vtvm at the diode (D201) cathode, and tune T202 for maximum reading. Next, couple a gdo or wavemeter to L206 and tune C218 for maximum. Then couple a signal at the operating frequency to the input of the front end. Peak L201, L203, L204 and L205 for maximum audio output. If Q201 oscillates, adjust neutralization coil L202. Turns may have to be removed or added, depending on the individual characteristics of Q201 and because of the limited tuning range of the coil.

The above has been a preliminary adjustment. A cooperating station on frequency can be used for final on-the-nose alignment. If an fm generator and sweep generator are available, by all



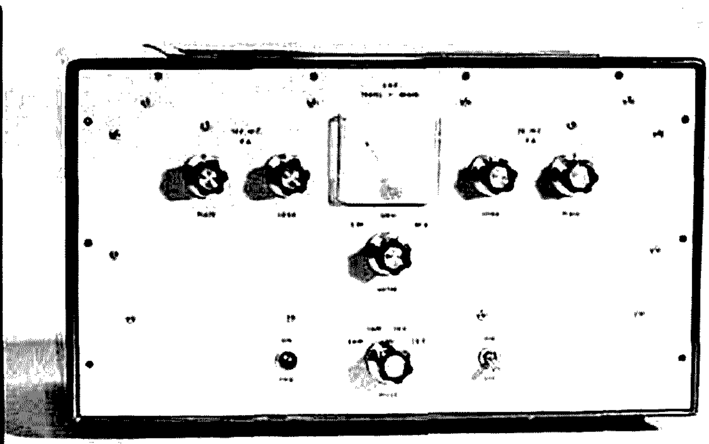

| | |
|---|---|
| D101 | 3.9 volt zener diode, 1 watt (1N4730) |
| L101,L102 | 5.6 mH (Nytronics SWD-5.6 or equivalent) |
| L103 | 82 mH (Nytronics SWD 82000 or equivalent) |
| T101 | CTC 3624-2 coil form wound full with no. 32, 3-turn link, shielded (winding area of CTC 3624 is 0.183" long, 0.210" winding diameter) |
| T102,T103 T104,T105 T106 | Miniature 445-kHz i-f transformer, 25k—600 ohms (J. W. Miller 9-C1 or equivalent) |
| T107,T108 T109 | CTC 3624-2 coil form wound full with no. 32, 8-turn link (see coil form info under T101) |

**fig. 2. I-f strip, discriminator, and af circuits. Bandpass filter following mixer consists of ordinary i-f transformers, allowing adjustable selectivity.**

means use them; however, an on-the-air signal is adequate. Peak all coils in the converter as well as in the selectivity strip (if used) and transformer T201 in the i-f strip for maximum audio output. If an on-the-air signal is used, attenuation may be necessary at the antenna input to keep the receiver from limiting. A weak signal is best. Transformers T102, 103, and 104 are stagger-tuned for desired bandwidth. Using a properly modulated on-the-air signal, tune these transformers for minimum distortion and clipping. T105 is adjusted for maximum audio output. T106 quadrature coil is adjusted for best audio quality. If a scope is connected to the audio output of U102, T106 should be adjusted for a symmetrical waveform. If more than one frequency is used, tune the rf coils for approximately the center frequency spread. All that's necessary now is to adjust the volume and squelch controls for desired levels.

One of the authors has been using this receiver for mobile operation daily for several months and has found that it performs quite well. The mobile transmitter is a hybrid affair, but work is under way on a 100-percent solid-state transmitter. We will present this to *ham radio* readers at a later date if sufficient interest is shown. There is no reason why anyone who wants to get on vhf fm today can't build a receiver such as we have presented and produce a unit that will perform to expectations at reasonable cost.

**ham radio**



# a multimode transmitter for six and two meters

PC-board mixers featured in an earlier article are combined with linear amplifiers and control circuits for improved operation

D. W. Bramer, K2ISP, 45 Thayer Road, Fairport, New York 14450

A previous article in ham radio featured transmitting mixers for the two- and six-meter bands.[1] Regular low-frequency equipment provided excitation for the six-meter unit, while a 50-MHz source provided drive for the two-meter mixer.

Although these converters produce low output power, they are sufficiently complete to form basic subassemblies for a medium-power dual-band vhf transmitter. This article shows how to combine these printed circuit assemblies with a bias supply, control circuit, and linear amplifiers. Construction techniques will depend on individual requirements, so I've only highlighted physical details; these are shown in the photos and sketches. Circuit details are shown in the schematic (fig. 1).

Operating modes can be selected with a single control:

1. 6-meter a-m, using a low-power 50-MHz exciter such as the Heath "Sixer."

2. 2-meter a-m, using the same exciter as above.

3. 6-meter ssb, using a regular low-frequency ssb transmitter or transceiver.

4. 2-meter ssb, using the same low-frequency ssb source as above.

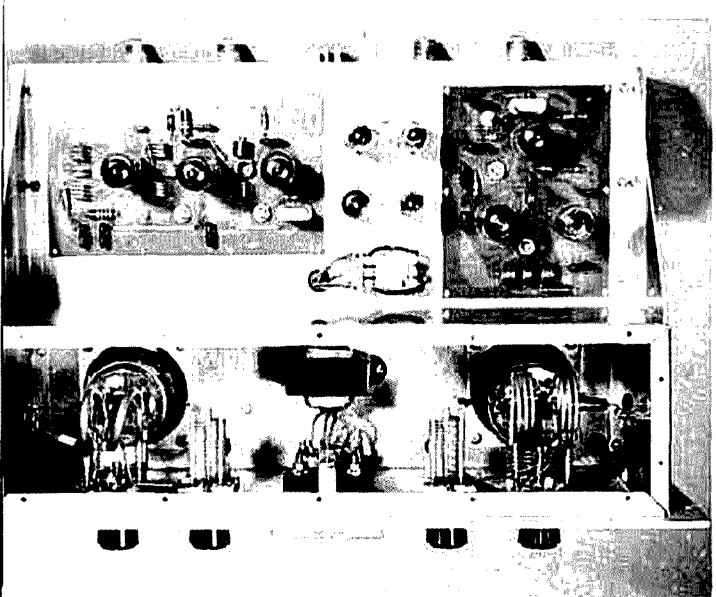
Carrier insertion allows cw operation in both mode switch positions. Break-in cw and vox are included to enhance enjoyment.

The Amperex 5894, which is interchangeable with the 829B, may be more efficient, particularly on two meters. However, if the 5894 is used some reduction will be required in neutralizing capacitance. Also, the grid-circuit inductance will have to be increased.

## metering system

With the selector switch in the center position, grid current of either amplifier can be monitored. In the counter clockwise position, 2-meter output power will be indicated; 6-meter output is indicated when the switch is in the clockwise position. No grid current should flow in Class $AB_1$. However, an indicator is required to show when grid current begins to flow to obtain maximum output with minimum distortion.

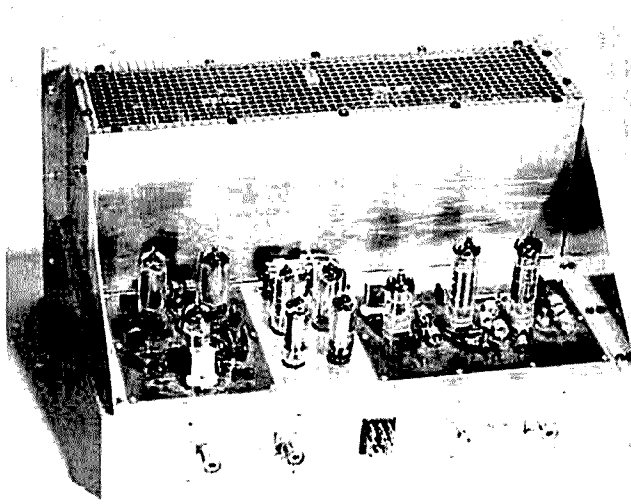
Chassis layout of the 6- and 2-meter multimode transmitter. The 2-meter transmitting converter is rear left; the 6-meter transmitting converter is right rear. The 829B final-amplifier stages are in the shield compartment to the front.



## power supplies

Regulated +210 volts are required for the 6U8 oscillator-buffer, 5763 screens, and 12AT7 plates. Regulated Screen voltage for the 829B's is also required. Two independent series strings of OB2 regulators are used. This provides isolation between the oscillator and 829B screen-current fluctuations. A simple bias supply is used. A small 6.3-volt transformer is connected backward in the filament line. Its unloaded output (−145 volts) is applied through two resistors to 18- and 22-volt zeners. When the control relay is unenergized, both diode supplies float, allowing the bias lines to assume full

Rear view of the multimode transmitter shows the neat layout and construction used by the author.



muting potential of -145 volts. In the energized position, the zener circuit is grounded, providing -18 and -22 volts bias for the mixers and 829B grids respectively.

Supply voltages are brought in via a 12-prong male Cinch-Jones plug mounted on the chassis rear apron. Supply requirements include +800 Vdc, 150 mA; +300 Vdc, 275 mA; and 6.3 Vac, 6.5 A.

## construction notes

My construction techniques are apparent in the photos. Three major structures

are used; an L-shaped main chassis plate, a U-shaped rf shield for the final amplifier compartment, and a 1/8-inch-thick front panel. Rough dimensions are shown in fig. 2. Angular side supports can be added, as shown, to strengthen the assembly. Except for the front panel, all pieces are made from 0.05-inch-thick 5052-H34

| | |
|---|---|
| C1, C4 | 30-pF trimmer (Arco 461) |
| C2 | 25-pF butterfly variable (E. F. Johnson 167-22) |
| C3 | 140-pF air variable (Hammerlund HF-140) |
| C5 | 10-pf butterfly variable (E. F. Johnson 167-21) |
| C6 | 100-pF air variable (Hammerlund HF-100) |
| L1 | 5 turns no. 16, 7/16" diameter |
| L2 | 6 turns no. 16 each side of center, 5/8" diameter |
| L3 | 3 turns no. 10 each side of center, 1¼" diameter |
| L4 | 4 turns no. 14, 7/8" diameter |
| L5 | 2 turns no. 16, ½" diameter |
| L6 | 3 turns no. 12, 5/8" diameter, 1½" long, center tapped |
| L7 | 2 turns 3/16" silver-plated tubing, 1¼" diameter, 1-1/8" long, center tapped |
| L8 | 1 turn loop around center of L7. Refer to photo for approximate size and position. |

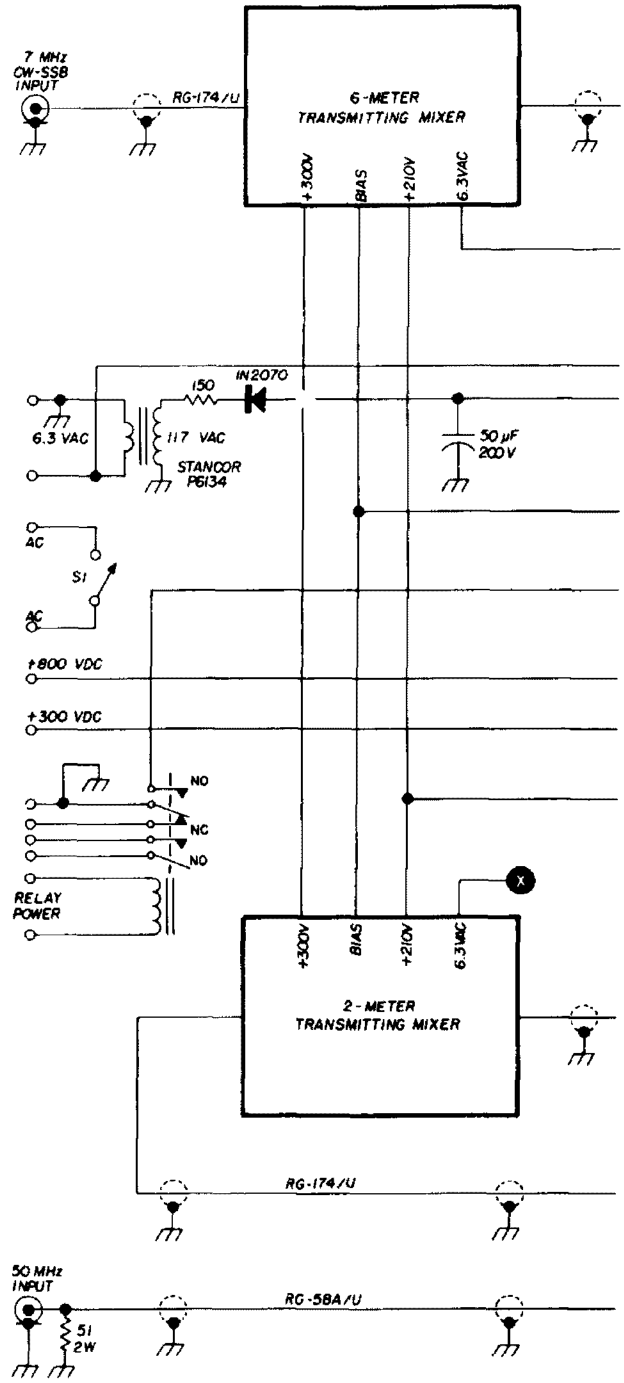
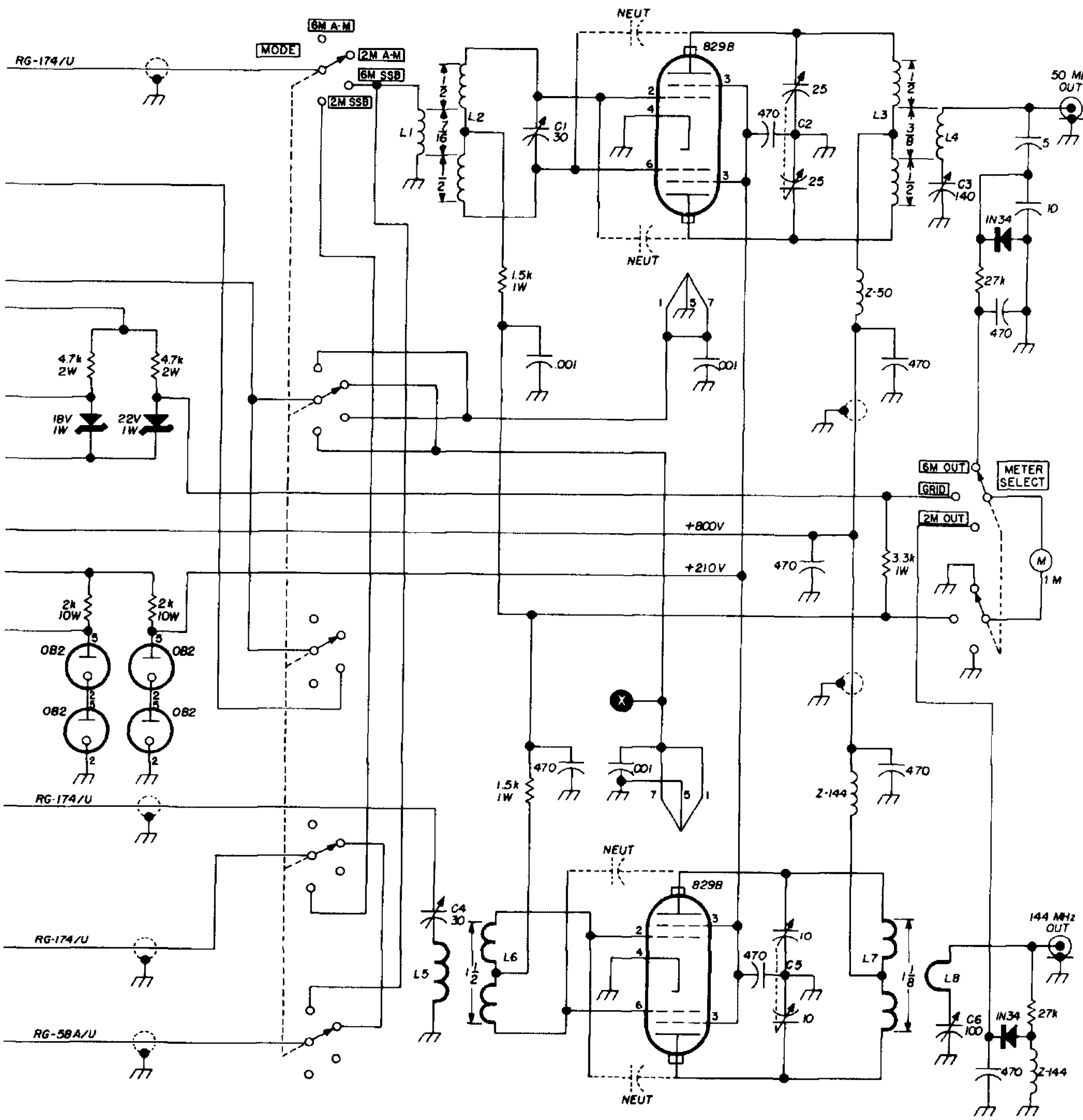



fig. 1. Schematic diagram of the 6- and 2-meter multimode transmitter.

aluminum. An aluminum angle, ½ x ½ inch, was drilled and tapped to form the fourth top surface edge of the rf shield compartment. I tailored the front panel to accomodate the assembly in a Heath "Seneca" cabinet, which I obtained from a surplus outlet. The knobs are from Heath.

Final amplifier components, panel meter, meter switch, and the bias supply transformer are mounted above deck inside the shield compartment. The control relay and voltage regulator tube sockets are mounted near the chassis rear center, outside the shield compartment.

The 829B sockets are E. F. Johnson type 122-105-100. The circular portions extend well below the chassis. Grid coils are suspended from the socket terminals with number 14 insulated solid wire leads. These are criss-crossed and extend up through ceramic wafer holes to form

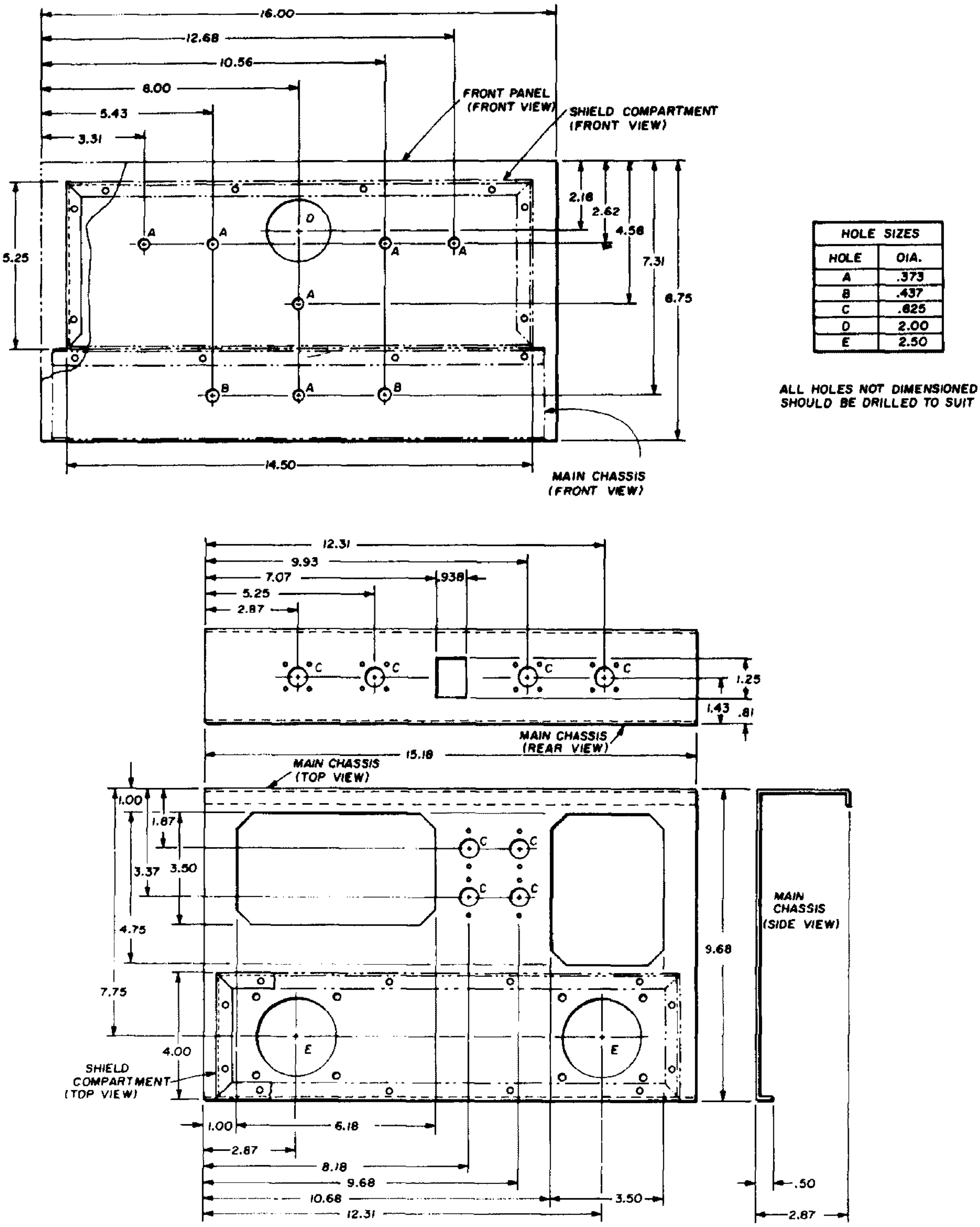



neutralizing capacitors. The leads should run about one-half inch above the chassis surface. Capacitance is adjusted by bending the leads.

Link input coils, 1.5 kilohm grid resistors, and the bypass capacitors are suspended from a small terminal strip soldered directly to the base of each socket between pins 1 and 7. High voltage for the 829B tank is brought above deck with RG-58/U coaxial cable. Fahenstock spring clips, soldered directly to butterfly variable capacitors, provide 829B plate pin connection.

## adjustment and operation

Reference 1 should be reviewed before attempting to set up this more complex system. If the printed-circuit subassemblies have been preadjusted, the remaining task is to stabilize the 829B stages and optimize their input coupling. Reference




| HOLE | DIA. |
|---|---|
| A | .373 |
| B | .437 |
| C | .625 |
| D | 2.00 |
| E | 2.50 |

**fig. 2. Mechanical construction details of the 6- and 2-meter transmitter chassis.**

to the ARRL Handbook will be helpful in obtaining correct adjustment.

First, apply heater power only. The appropriate heaters should light with the mode switch in each position. Next, check for -145 volts on the 829B grid and mixer-board lines. With the relay armature depressed, -22 and -18 volts respectively should appear on 829B grid and mixer lines.

## two-meter a-m adjustment

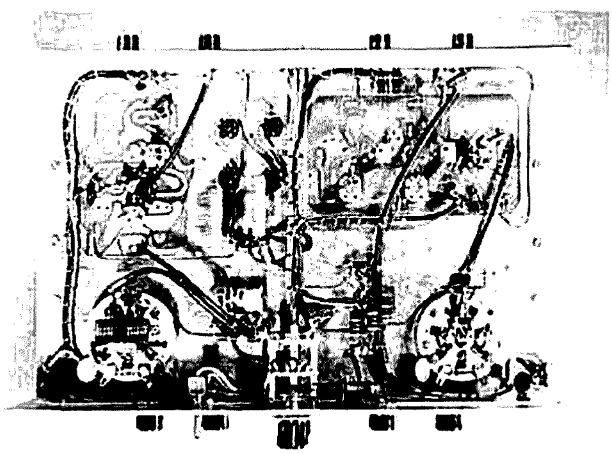
Connect a 50-ohm dummy load to the 50-MHz output jack. Position the mode switch to "6 AM." With heater power and drive applied, the 829B stage can now be

neutralized. (A standard procedure is given in the ARRL Handbook.)

Apply 300- and 800-volt dc power. (It might be helpful to reduce the 800-volt power to 500 volts for this initial adjustment.) Adjust the controls for maximum power into the dummy load. This can be observed on the panel meter with the meter switch in the "6M OUTPUT" position. Peak trimmer capacitor C1 for maximum output indication on the meter.

## two-meter a-m adjustment

Connect the 50-ohm dummy load to the 144-MHz output jack. Leave the 50-MHz exciter connected. Rotate the mode selector switch to the "2 AM" position and the meter switch to "2M OUTPUT." Neutralize the 829B stage. Apply dc power and tune for maximum rf power into the dummy load. Alternately adjust capacitor C4 and the two-meter mixer board output capacitor (butterfly variable) for maximum 144-MHz output.

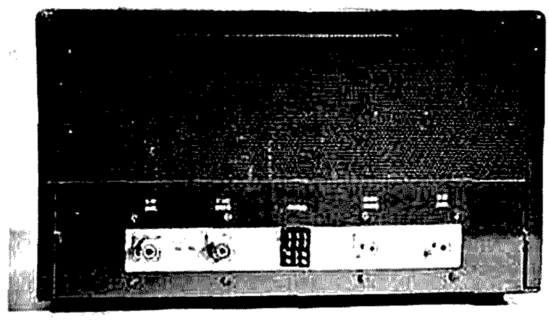


**Under-chassis shows good vhf construction.**

## six- and two-meter adjustment

Connect suitable excitation to the low-frequency input jack. Rotate the mode switch to the "6 SSB" position. With the meter switched to read "6M OUTPUT," gradually increase drive for 50-MHz output indication. Peak the 6-meter mixer-board output tank butterfly variable capacitor for maximum indication. Position the meter switch to read grid current, and increase drive until an indication of grid current is just apparent. This represents the maximum drive level at which maximum output will occur consistent with minimum distortion.

The unit is now adjusted for the "2M SSB" operating mode. Now rotate the mode switch to the extreme clockwise position, and adjust the controls for maximum output while observing the drive level.



**Rear panel of the multimode transmitter.**

## final notes

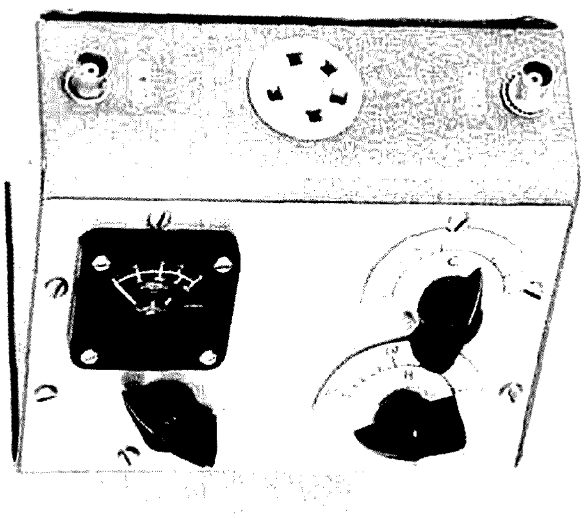
It may be necessary to alter the value of the resistors in the output sampling circuit. Try different values near 27 kilohms until a good meter deflection is obtained at maximum output.

For a-m operation, tune for maximum output in the ssb mode first, then switch to the a-m mode position. The 50-MHz a-m input source must be attenuated to produce about one-half the ssb-mode output as indicated on the meter. Modulation up to 100 percent will then be amplified linearly. Peak envelope ssb output is about 50 watts, while output on a-m is about 12 watts on either band. Third-order distortion products on 144 MHz measure about 30 dB down.

### reference

1. D. W. Bramer, K2ISP, "Heterodyne Transmitting Mixers for Six and Two Meters," *ham radio*, April, 1969, p. 8.

ham radio



# a simple bridge for antenna measurements

This instrument allows measurement of both resistive and reactive components of your antenna with the aid of the Smith chart

Henry S. Keen, W2CTK, 64 Schuyler Drive, Commack, New York 11725

Articles published in the amateur literature on the subject of the Smith chart[1] have been of more academic interest than practical value. How many hams, for example, have slotted-line and swr-indicator facilities appropriate to the frequencies involved? The little device described in this article can be built by anyone and will provide useful data for antenna measurements based on the Smith chart, even for the "dc bands."

## basic bridge circuit

The need for more accurate evaluation of a recently erected 15-meter quad was indicated when a borrowed vswr meter provided a value that was obviously too good to be true. The bridge described here was the outcome and has provided much useful information.

The basic bridge circuit (fig. 1) will measure a purely resistive load with acceptable accuracy; but when a reactive component is present, the null will not only be broad, it will be displaced and therefore inaccurate.

Several methods are available for balancing and measuring the reactive component, but the simplest appears to be

placing a parallel-resonant tuned circuit across the load, as shown in **fig. 2**. By detuning this circuit from resonance, it's possible to introduce a controlled opposite reactance across the load and adjust the bridge to a perfect null. The reactive component of the load is measured by the direction and degree of this detuning. The equivalent circuit of the antenna therefore is represented by a resistance shunted by a reactance. Because these elements are in parallel, it's probably more accurate to regard this bridge as an admittance rather than an impedance bridge, and to perform any computations on that basis.

Two arms of the bridge are made up by the 100-ohm linear potentiometer. A third arm is a 51-ohm, half-watt carbon resistor, while the fourth arm is the load, shunted by the parallel-tuned circuit.

The variable capacitor should be at least 250 pF for most applications, while the switched coil sections (see photo) are merely sufficient in number that resonance on each band can be obtained on at least two settings of the selector switch. This provides for adjustment in either direction from resonance when a complex load is present.

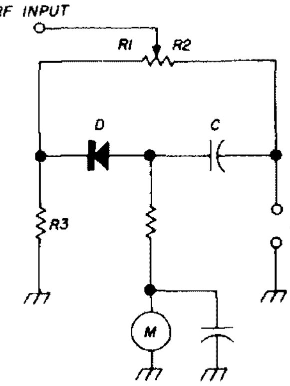



**fig. 1. The basic rf bridge. At balance, R1/R2 = R3/RX.**

A germanium rather than a silicon diode was selected because of its lower barrier potential. Although a 100-$\mu$A meter was used, its choice was dictated more by small size than current rating. For greater sensitivity, the 1.5k isolating resistor can be replaced by an rf choke.

## calibration

No measuring device is better than its calibration. Many hams may feel they have too little precision equipment with which to calibrate a device of this type. However, a nice thing about this bridge is that the resistive and reactive controls can be calibrated separately. Before wiring the bridge, the resistive control can be calibrated against a purely resistive load. The resistive calibration can even be made with a couple of flashlight cells using the circuit of **fig. 3**. A handful of 100-ohm, half-watt resistors used in various combinations should provide sufficient useful calibration points. If you use as many resistors as possible (within reason of course) in these combinations, individual differences will tend to cancel.

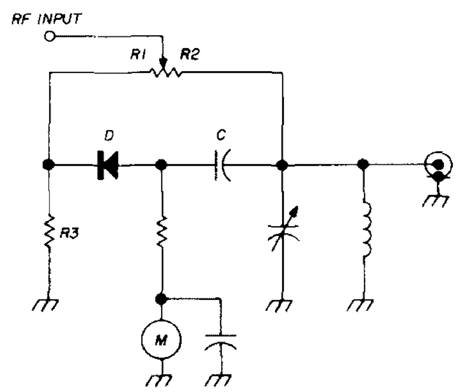



**fig. 2. Modification to measure complex impedances.**

Calibration of the capacitor is greatly simplified if you have access to a Q meter. Circuit capacitance, and the minimum capacitance of the variable capacitor itself can be ignored, as we are interested only in the incremental accuracy; i.e., the difference in capacitance between any two settings of the dial. A handful of silver-mica capacitors of known value can be used with a grid dipper for the calibration, ignoring any points that don't lie on a smooth curve. A preliminary calibration by this substitution method was later repeated, using a precision decade-resistor box and a Q

meter. The difference between the before-and-after measurements were well within 10%, which is a good working value around a ham station.

## measurement procedure

A necessary device for this application is a reasonably well-matched 50-ohm load. A half-watt, 51-ohm resistor, mounted in a BNC connector performed quite well.

At a given frequency, the rf input to the bridge is first adjusted to produce full-scale deflection of the meter with the controls set at either extreme. The bridge is then balanced at this power level against the reference load, and the capacitor reading is recorded. The antenna is then substituted for the reference load, and the bridge is rebalanced. These resistive and capacitive readings are recorded, and the process repeated on other frequencies of interest.

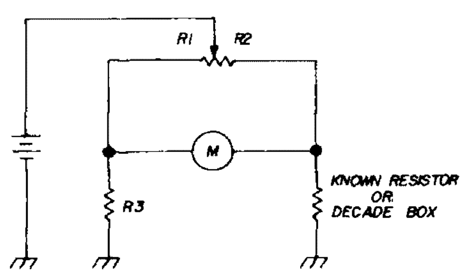



**fig. 3. Calibration circuit for the resistive control.**

To compute the equivalent antenna impedance, (or admittance) a table is set up as in table 1. For each frequency the setting of the capacitor that was obtained with the reference load in place is listed, then the resistive and capacitive readings with the antenna in the circuit. The two capacitive readings for each frequency are compared and their differences listed. The reactance, which must be computed for each frequency (represented by this differential capacitance) is then listed in the next column.

Normalized conductance and susceptance are next computed for a 50-ohm reference level by dividing 50 by the measured resistance and the computed reactance of the capacitive differential. If you're using 70-ohm line, you should substitute 70 for 50; otherwise all procedures are the same. If the setting of the capacitor was reduced by substitution of the antenna for the reference load, the susceptance is positive (capacitive). If it was increased, the susceptance is negative (inductive).

**Table 1. Data for computing equivalent antenna impedances.**

| Freq. (MHz) | R (ohms) | C (pF) | Ref C (pF) | $\Delta$C (pF) | X | G' | B' |
|---|---|---|---|---|---|---|---|
| 21.0 | 60 | 20 | 75 | 55 | 137.6 | .833 | .363 |
| 21.1 | 61 | 26 | 72 | 46 | 163.8 | .820 | .325 |
| 21.2 | 64 | 30 | 73 | 43 | 175.0 | .780 | .285 |
| 21.3 | 68 | 29 | 72 | 43 | 174.0 | .737 | .287 |
| 21.4 | 76 | 24 | 71 | 47 | 158.0 | .659 | .315 |
| 21.5 | 81 | 14 | 70 | 56 | 132.0 | .618 | .379 |

These normalized admittances can now be plotted on a Smith chart, as in **fig. 4**, and successive frequency points connected by a smooth curve. This curve showed my quad to be resonant at about 21.15 MHz. As I normally operate ssb at about 21.35 MHz, the resonant point represented about 4 or 5 inches of wire in the length of the driven element.

## smith chart

To convert your admittance measurements to more familiar impedance values, all that's necessary is to replot each point on the Smith chart at the same distance from the center of the chart, but located diametrically opposite the admittance

points. When making this transformation, bear in mind that when we deal in admittances the equivalent circuit is a resistive and reactive element in parallel; while the *equivalent circuit* for impedance is a resistive and reactive element in series. To return to absolute values from the normalized quantities on the Smith chart, merely multiply the impedance components by 50 or 70, as the case may be.

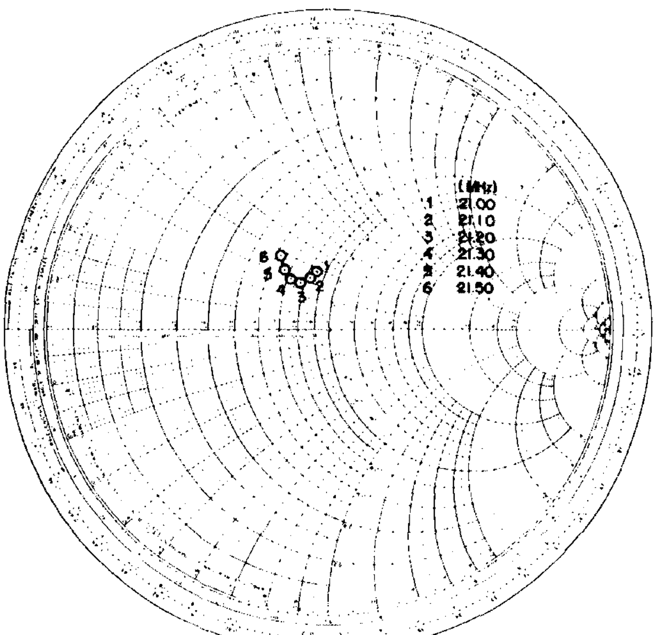



**fig. 4. Admittance plot of 21-MHz quad as measured through the 50-ohm coaxial line.**

The distance of the plotted point from the center of the chart **(fig. 4)** indicated that the vswr at resonance, (the points closest to the center, in this case) is about 1.5.* Although this is a reasonable figure, the match can be improved in several ways. First, and most practical, is to replace the 50-ohm cable with 70-ohm cable. Second, the spacing between antenna reflector and driven element can be reduced until the antenna input impedance more closely approximates 50 ohms; and third, a matching network of some kind can be installed at the antenna end of the coaxial line.

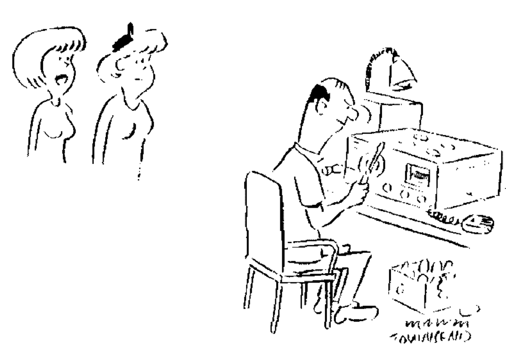
Occasionally one hears of someone who trimmed the length of his coaxial line "to make it load." What probably happened was that the antenna was so badly mismatched that, coupled with an odd length of coaxial line, a shunt reactance appeared across the output network that was so large the network just couldn't compensate for it. Trimming the coaxial line didn't reduce the vswr, but it did reduce the reactive component to where the overworked matching network was finally able to handle the situation. If you're forced to this quick and dirty solution, at least the Smith chart will tell you exactly how much transmission line to remove. In **fig. 4** it can be seen that *moving* the impedance plot about .092 wavelength nearer the load should remove most of the reactive component.

## transmission-line length

A fair representation of what the antenna looks like to the feed line may be obtained by transferring each admittance (or impedance) point toward the load (counter clockwise) a distance equal to the electrical length of the feed line in wavelengths at each frequency. Each half-wavelength represents a complete revolution around the center of the Smith chart. This operation assumes the line to be lossless, frequently an optimistic assumption!

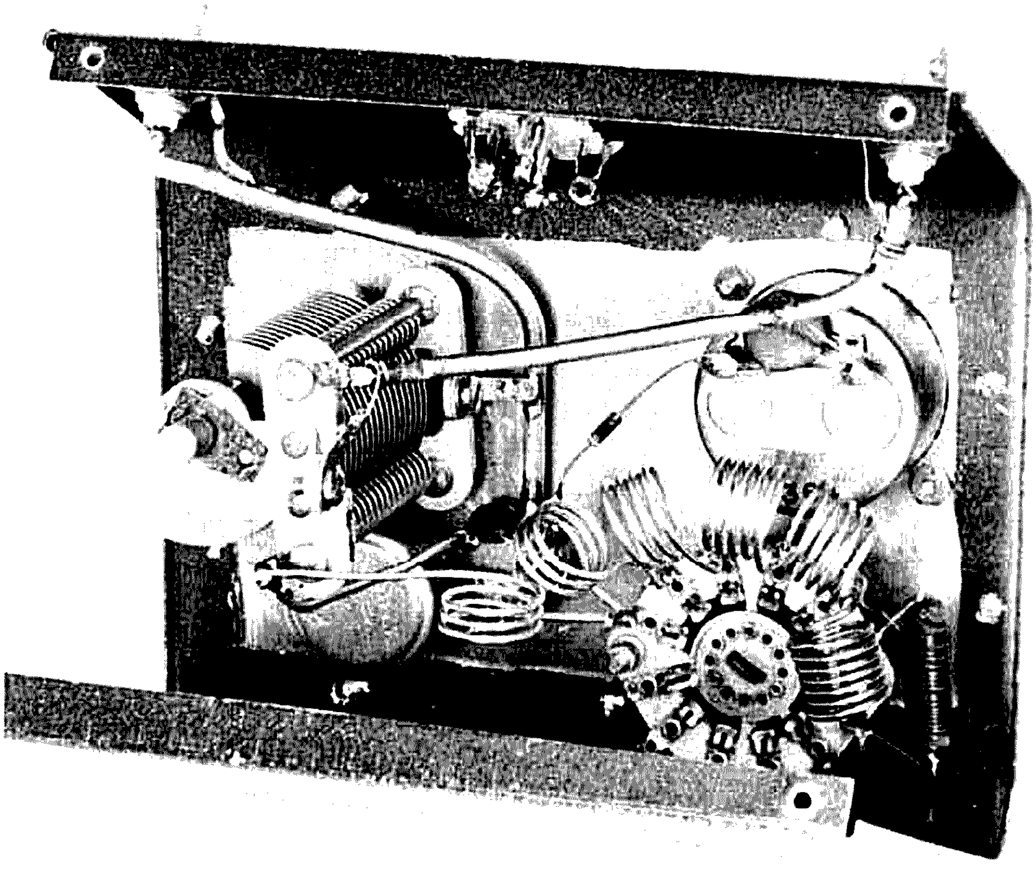
However for this move, the exact electrical length of the transmission line in wavelengths must be known. One method is to disconnect the center con-

*Because the circle through the resonant point intersects the $R/Z_0$ axis on the right-hand side of the chart at $R/Z_0 = 1.5$. **Editor.**



**"George is one of the silent majority . . . He can't get his transmitter to work."**

ductor at the antenna and grid-dip the line from the bottom end, coupling to a small one-turn coil. Above all don't trust the frequency calibration of the grid dipper, but check with a good wave-meter or calibrated receiver. If the transferred measurement points don't center on the scaled diameter of the Smith chart, your electrical length was inaccurate. Grid dipping to the odd harmonic nearest the operating frequency is probably best, but be sure you know which harmonic it is. The electrical length of the feed line is proportional to the frequency.



Inside the admittance bridge. Switched coil sections resonate on each band of interest.

## conclusion

Transferring the admittance or impedance data up to the business end of the feedline in this way will give you a piece of information otherwise hard to come by. Suppose, for example, an antenna shows a vswr of 2:1. You can't always be sure the antenna looked like 100 ohms, for it could as well appear as 25 ohms to the 50-ohm cable. You'll therefore be able to tell in which direction the loading adjustment should be made.

Although this type of antenna measurement may seem to involve considerable pencil pushing, it results in much better than a ball-park estimate. Those microvolts you're able to put into the other follow's receiver are expensive and hard to come by. It's nice to feel that you're getting everything possible from your installation because you did the job thoroughly, rather than settle for the too-frequent lick and a promise of which so many are often guilty.

### reference


1. F. E. Terman, "Electronic and Radio Engineering," 4th ed., McGraw-Hill, New York, p. 100.


ham radio

# neutralizing small-signal amplifiers

## How to obtain maximum performance from preamps and converters

Earnest A. Franke, WA4WDK, 108 Matawan Terrace, Matawan, New Jersey 07747

**In most of the converters** and preamplifiers built today the tubes or transistors are operated in the grounded-cathode or grounded-source configuration. Neutralization of the amplifier is necessary to compensate for the interelectrode capacitance in the tube or transistor. If not neutralized, this capacitance allows energy developed in the output circuit to be fed back to the input. The amplifier then acts as a tptg oscillator.

Amateurs unfamiliar with the theory of this phenomenon frequently encounter instability problems and find amplifiers hard to neutralize. This article provides basic data on neutralization and gives some hints on how to solve this problem. A properly neutralized amplifier will reward you with maximum gain and stability consistent with minimum noise figure, which is especially important at the higher frequencies.

### gain measurement

Amplifier gain can be measured in the usual manner, from input to output. Gain in the reverse direction can be measured also, and should be zero or nearly so. **Fig. 1** illustrates this concept.

The interelectrode capacitance in the amplifier allows a signal to be fed back from output to input. As the feedback capacitance is increased, the reverse gain also increases. Thus it is more important to neutralize amplifiers at the higher frequencies because of the decreasing capacitive reactance.

### neutralizing methods

When neutralization is required over a small bandwidth (as in an i-f amplifier), a resonant system can be used. An inductor, $L_n$, is made to resonate with the interelectrode capacitance and any shunt capacitance existing in the circuit **(fig. 2)**. The inductance value can be approximated by

$$L = \frac{1}{4\pi^2 f^2 C}$$

where f is the mid operating frequency, and C is the grid-plate or gate-drain capacitance of the tube or transistor. A large blocking capacitor, C, is inserted in series with the inductor to prevent shorting the supply voltage to ground through the input circuit. This blocking capacitance must be large enough to offer a negligible reactance at the operating frequency. Button capacitors are excellent because of their short lead length. Air-wound inductors are preferred for the neutralization coil, because they have less stray capacitance than the slug-tuned type. The neutralization point may be found by using a tuning wand with a brass or iron slug at each end. Proximity of the iron slug increases the neutralizing coil inductance, requiring the coil

to be squeezed. Brass decreases the inductance, requiring the coil to be expanded to reach resonance.

At resonance the impedance of a parallel circuit is a resistance that is Q times the reactance of either the inductance or capacitance.[1] Very high impedances can be developed by parallel resonance. Therefore, a higher-Q neutralization circuit will result in a larger isolation, as seen in **fig. 3**. As gain or bandwidth increases, the Q or impedance of the inductance must be improved to isolate the input and output circuits. Shunt capacity to ground must be avoided, which reduces the gain-bandwidth product.[2]

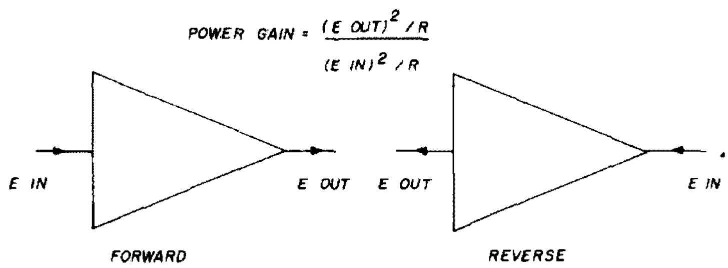



**fig. 1. Amplifier gain can be measured in forward or reverse direction. In the equation for power gain, R is the tuned-circuit impedance.**

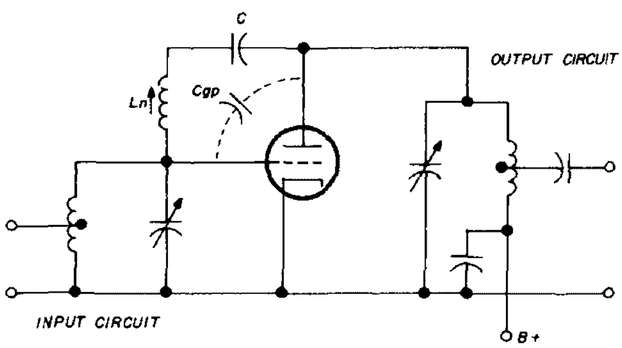



**fig. 2. Neutralization scheme for an amplifier operating over a very narrow bandwidth or at a fixed frequency. This is known as the *coil* neutralizing method.**

Neutralization is necessary to reduce tuning interaction between input and output. Many amateurs simply tune the neutralization coil in the amplifier until the output circuit doesn't *affect* the input. When the "blurps and squeals" disappear, neutralization is considered to be complete.[3]

## tube neutralization

When tubes are used, neutralization is simplified because the plate voltage is removed. The neutralizing coil is adjusted for a strong signal applied to the input. At the minimum output point, the coil resonates with the grid-plate capacitance, and circuit impedance is maximum. A simple neutralization procedure can be used, requiring a strong local signal to be fed to the input and a detector coupled to the output.

## fet neutralization

Field effect transistors are being used more today because of their low noise figure, ability to accept large signals with small cross modulation, and low cost. Inside the fet is a capacitance, $C_{rss}$: the common-source, short-circuited, reverse-transfer capacitance. This capacitance between the gate and drain is a parallel combination of a voltage-sensitive depletion capacitance and lead capacitance. The behavior of the depletion capacitance is similar to the variable-capacitance effect obtained by reverse biasing semi-

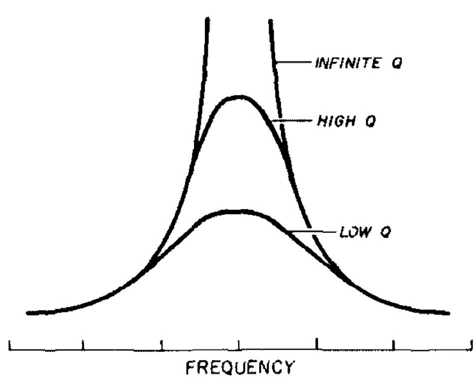



**fig. 3. Impedance as a function of frequency with circuit Q as a parameter (parallel-resonant).**

conductors.

Differing from the tube method, attempts to neutralize the fet by minimizing a signal applied to the input without the drain supply voltage applied will not work. The junction capacitance changes significantly when the supply voltage is applied.

## mosfet and grounded-gate circuits

Neutralization usually may be avoided by operating the amplifier in the common-gate mode. The grounded gate acts as a radio frequency shield between input and output. Feedback capacitance is no longer the gate-drain capacitance, but is the source-drain capacitance.

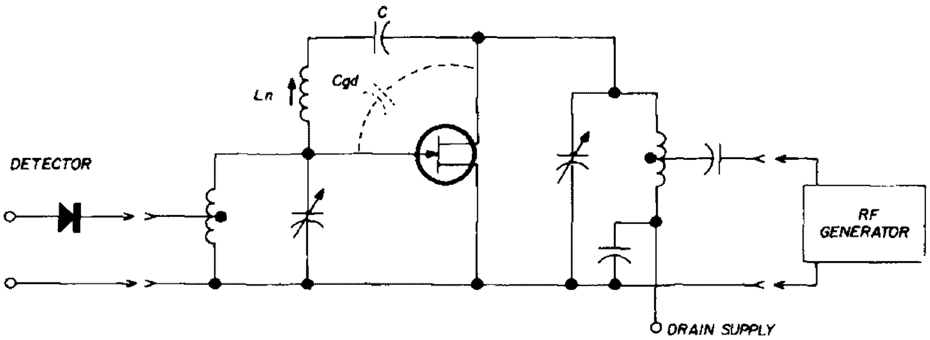



**fig. 4. Method of measuring reverse gain in a typical amplifier.**

## neutralizing with applied voltage

The following method can be used to neutralize a tube or fet amplifier with the supply voltage applied. An rf signal generator or sweep oscillator is coupled to the output circuit, as shown in fig. 4. Generator level is increased until an output is noted by a sensitive detector or receiver connected to the input circuit. The neutralizing coil, $L_n$, is tuned for a minimum indication at the detector. The amplifier is "analyzed backwards" in this method, and reverse gain is measured. If a sweep generator is used, the oscilloscope presentation would resemble fig. 5. A "suck out" will occur at the neutralized frequency,. caused by the isolating high-impedance resonant circuit. The bandwidth of the "suck out" will be determined by the Q of the resonant neutralizing circuit and must be wide enough to cover the amplifying bandwidth of the amplifier in the forward direction.

Complete neutralization may still not be achieved if stray inductive or capacitive coupling exists within the amplifier components or extends to another stage. The answer to this problem is shielding or rearrangement of components. Proper bypassing, with short lead lengths, must be maintained to avoid oscillation.

With the mosfet semiconductor, any necessary neutralization is largely due to socket-lead capacitance. The feedback in the common-source configuration is less than that in a vacuum tube in the grounded-grid mode. You might get by

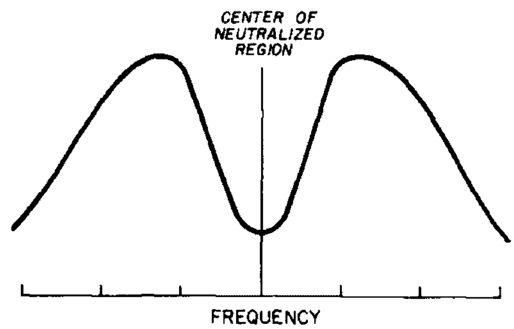



**fig. 5. Oscilloscope display of "suck-out" at neutralization frequency.**

without neutralization and still have a stable amplifier, but the noise figure and gain will be improved with it. Once tried, it is a simple procedure.

### references


1. Frederick Terman, *Electronics and Radio Engineering,* McGraw-Hill, p. 52.
2. Ibid., p. 415.
3. D. D. DeMaw, "FET Converters For 6 and 2 Meters," *QST,* May, 1967, p. 11.


**ham radio**

# electronic counter dials

A direct frequency readout for your receiver vfo using inexpensive ICs

E. H. Conklin, K6KA, Box 1, La Canada, California 91011

**One evening at the home** of Stan Dixon, VK3TE, several of the VK3 gang were chatting about things of general interest when Harold Hepburn, VK3AFQ, mentioned that dials for accurate frequency indication were difficult to obtain in Australia. He was building a two-meter receiver using a 10–10.5-MHz vfo and wanted something reasonably accurate for frequency readout. He asked about the possibility of using a built-in electronic counter for the vfo that would indicate frequency directly.

This set me thinking about the problem. After considering the many angles, I concluded that this would be a feasible project, using inexpensive digital ICs. This was before any details of the *signal/one* receiver were available, which uses a digital dial readout that gives calibration accuracy of 100 Hz on each band.

## conversion-frequency correction

All equipment won't necessarily have the convenient vfo range that VK3AFQ had in mind, but may begin at some odd frequency. Furthermore, heterodyne conversion oscillators will have some error (unless they're adjusted to exact frequency), which will be part of the readout.

This error may be different for different bands. Computer techniques could be used to remember the error and add or subtract the correction for the conversion frequency for each band, but a much simpler method is at hand.

If the bandswitch is provided with suitable contacts, the flip-flops in the frequency counter can be preset so that the counter will show the correct happening at the right time.

Some vfo's may tune up on some bands and down on others. This sounds like the last word in complexity for an electronic counter dial, but it's not impossible. By using gates that switch the circuit to count up or down, a single external contact can make the counter go either way. Several such circuits appear in references 2 and 3.

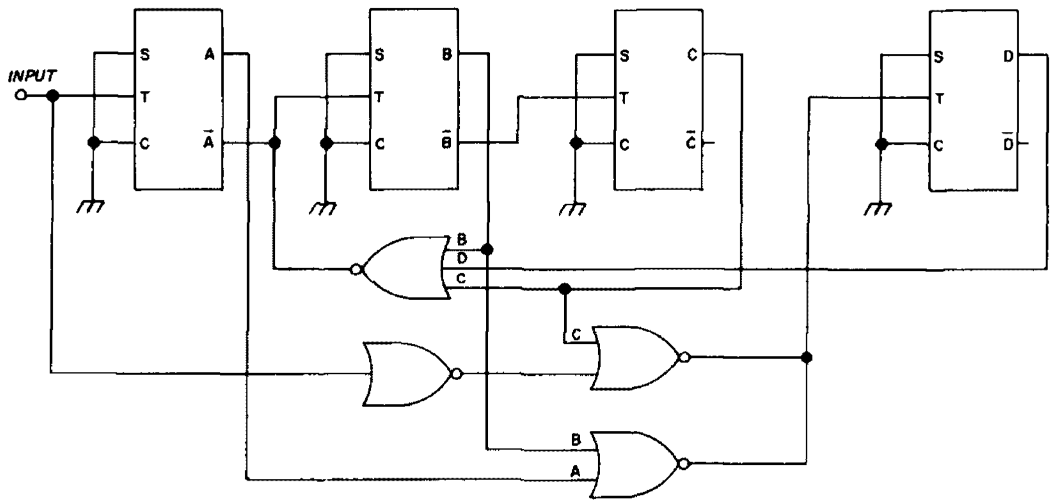



**fig. 1. A decade divider that counts down instead of up. Gating is such that, on leaving zero state, the circuit returns to the 9 state directly by inhibiting the B flip-flops and causing the D flip-flop to toggle during this condition only.**

frequency at the end of the dial. It would be necessary only to note the readout at the end of the tuning range, then wire the *preset* terminals of the counter ICs so that the additional count at the end-of-band vfo frequency would set the indicator to zero (000.0 kHz).

## backward vfo's

Many vfo's tune backward: the highest vfo frequency is the lowest receiver frequency. Most counters count up, not down. It would be useful to avoid a computer subtraction process. Motorola's Application Note AN-251[1] states, " . . . a ripple counter will count down when the complemented output from one stage drives the clock input of the following stage. . ." Motorola's circuit showing how this is done is reproduced in **fig. 1**. Four NOR gates are added to keep events

## time-base generator

A time-base generator will be required, as in other counters. If you would like to have an accurate frequency standard[4,5] it would be worthwhile to combine functions and use this accurate base for the electronic counter dial as well. However, if you wish to settle for a dial that displays only four digits (three for integral kHz and one for tenths of a kHz), some inaccuracy can be accepted. Two decades of the counter won't have any readout. You could use the 60-Hz power-line frequency to generate the time base. The time-base generator may require only four dual JK flip-flops for a one-second count. Hewlett-Packard, in the instruction book for their little four-digit counter, states that an accuracy of 0.02 percent or better is expected from the use of the power frequency for

gating.

I have completed experimental work using the power-line frequency for gating. A report on this will be in a forthcoming article. It confirms that the inaccuracies, which are within 0.02 percent, will not affect a counter dial with two digits not displayed. It appears that you may expect an error of one or two counts (tenths of a kHz) occasionally in a four-digit dial indicator readout that displays kHz and tenths of a kHz in four digits. This should be good enough for tuning indicators.

## display methods

It's possible to count vfo frequency, store the count, and display only the last completed count until it must be changed. (See the data on the Fairchild CL 9959 buffer-storage element and the CL 9960 decimal-decoder driver.) The count can remain unchanged until some later count requires the readout indicator to change to a new frequency. However, although storage, decoding, and lamp-driver ICs are available, we can get along with something even simpler.

Let's say we can use some of the available percent error to reduce the count period to 0.1 second rather than the more common 1-second interval. Without storing, or blanking the indicator during the count, we would then have 0.1 second during which the indicator runs through a count and gives no clear indication of frequency. This can be followed by nearly 0.9 second during which an unchanged frequency is displayed, even when the dial is being turned.

This action can be accomplished by using some of the ICs to produce a 0.1-second gating input, the total 1-second display period, and a preset signal before the start of the next count. As indicated earlier, this can preset the combined heterodyning frequencies for one end of the tuning dial, so that when the vfo frequency is counted at this point, the indicator will show 000.0 kHz.

The indicator can be any of the several types of digital displays. The less-expensive displays, such as the gas-filled National Electronics NL-950 shown in the Newark and Allied catalogs, cost about $6.25 per digit, or $25.00 for the entire four-digit display. For those who are satisfied with the binary-digital combination[5] and are willing to read each digit in an 8-4-2-1 combination using four lamps, a somewhat lower cost is possible.

## conclusion

Now let's see what we have. Inside our set will be a PC board about 4 by 6 inches, plus some space for the decoding gates or lamp-driver transistors. The vfo bandswitch will have contacts that preset the counter ICs to the correct frequency for the end of the vfo tuning range.

Outside, the unit will have a suitable knob with a convenient tuning ratio, but no associated indicator. Above it will be either four indicating gas-filled tubes or four rows of four lamps each for binary readout.

When the unit is turned on, the error or drift in the crystal-controlled heterodyne conversion oscillator won't be corrected in the tuning readout, but vfo error will be. If the knob is turned rapidly, the indicator will show an approximate frequency at 1-second intervals, thus lagging a bit behind a rapidly turned knob. When the knob rests for about a second, however, the readout will show the frequency to an accuracy close to 0.1 kHz.

### references


1. "Application Note AN-251," Technical Information Center, Motorola Semiconductor Products, Inc., P. O. Box 13408, Phoenix, Arizona 85306.
2. "DCL Series 800 Handbook," May, 1968, Signetics Corporation, 811 E. Arques Ave., Sunnyvale, California 94086.
3. "Signetics Utilogic II Handbook," Signetics Corporation, 811 E. Arques Ave., Sunnyvale, California 94086.
4. E. H. Conklin, K6KA, "Amateur Frequency Measurements," *ham radio,* October, 1968, p. 53.
5. E. H. Conklin, K6KA, "Frequency Counters and Calibrators," *ham radio* November, 1968, p. 41.


**ham radio**

# solid-state audio oscillator-monitor

The basic for this efficient, low-power circuit is an isolated integrator network

N. J. Nicosia, WA1JSM, 85 North Street, North Reading, Massachusetts 01864

**Prompted by a recent** article in *ham radio,*[1] I fulfilled a long-time desire to design an audio oscillator that works with digital integrated circuit supply voltages, produces a clean sine wave at 1 kHz, drives a speaker, and is inexpensive and easy to build. The result is described in the following paragraphs.

## circuit description

A minimum number of components is used in an efficient sine-wave oscillator circuit **(fig. 2)**. Transistors Q1 and Q2 form a high input and low output impedance amplifier, a feature of operational amplifiers. A "basic isolated integrator network"[2] is inserted between input and output. The circuit can be made to

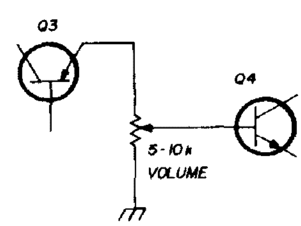



*fig. 1. Optional circuit for volume control.*

*oscillate at the frequency determined* by the network constants by omitting a resistor between the base of Q1 and ground. Slight adjustment in frequency be prevented from heating up by adding more resistance in series with the speaker. I don't use a volume control, since the level of my unit is perfect for my

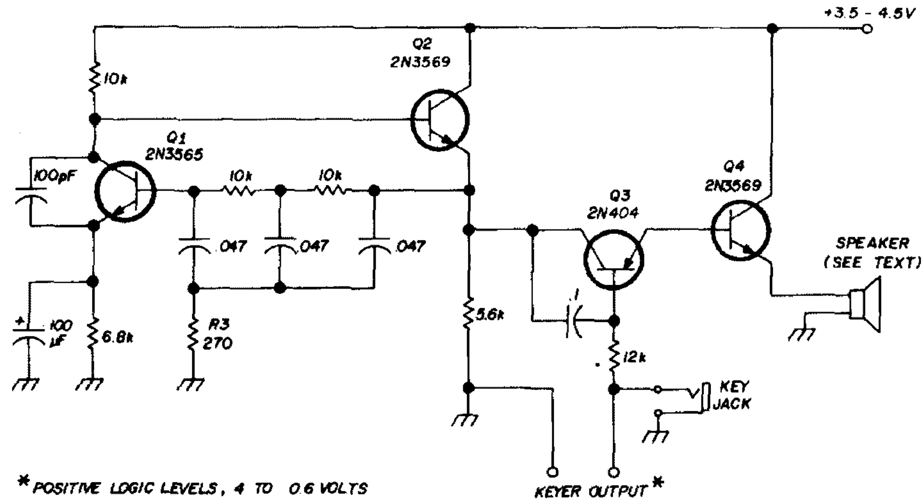



**fig. 2. Schematic of the 1-kHz audio oscillator. Output frequency is determined by integrator circuit between Q1 and Q2.**

can be made by changing the value of R3; however, don't make too big a change since stability may be affected. Q3 operates as a series switch and is turned on by grounding its base resistor. The transistors specified for Q3 are germanium types because of their low saturation voltage, which promotes efficient switching. Q4 is an audio power amplifier and drives a speaker directly.

I use a 3.2-ohm speaker in series with a 22-ohm resistor as a load. Any speaker will work, even the 40-ohm transistor-radio types. Use a scope to ensure a good waveform across the whole load. Q4 can requirements. However, if one is desired, see **fig. 1**.

## power supply

With a 4-volt supply and key down, the total power drain is 35 mA. For a power supply you *can use three* D cells in series or obtain 3.6 to 5 volts from your electronic keyer. A simple power supply is shown in **fig. 3**.

Logic levels of 4 to 0.6 volts are commonly encountered with keyers. I would be interested in hearing how the circuit can be tied into the many keyers described in the amateur literature.

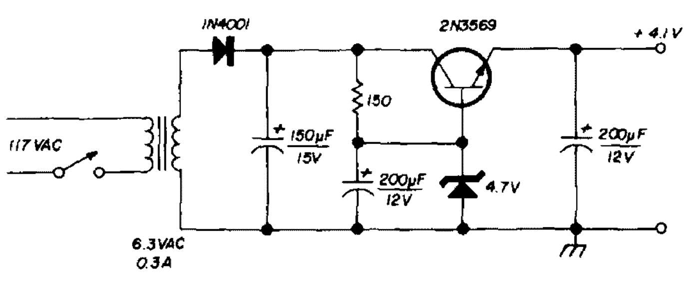



**fig. 3. Suggested regulated power supply.**

### references


1. G. D. Young, VE7BFK, "Electronic Keyer," *ham radio*, November, 1969, p. 32.
2. Ralph Glasgal, "Tunable RC Null Networks," *E.E.E.*, October, 1969.
3. Paul V. Wanek, "Nomograph Charts—A Fast Way To Build A Notch Filter," *Electronics*, September, 1969.


**ham radio**

# nomograph for reactance problems

An easy-to-use computational tool that lets you solve complex inductive and capacitive reactance problems without mathematics

Alf Wilson, W6NIF, 3928 Alameda Drive, San Diego, California 92103

**Nomographs are aids for quickly** solving many electronic circuit problems. A straight-edge placed across appropriate scales allows you to solve equations without using a slide rule or pencil and paper.

The nomograph in **fig. 1*** can be used to solve reactance problems when one quantity is unknown and two are known. Chart A is used to determine magnitude and decimal-point location. The significant figures are determined from chart B.

## practical example

Suppose you're interested in a circuit such as that shown in **fig. 2**. It's a Q-multiplier that can be added to your strip for increased selectivity. Let's say you have an inductance of fairly high Q whose value is 5 mH. You'd like the circuit to resonate at 1 kHz; what value capacitor should you use in the op amp feedback circuit?

You could determine the capacitor's value by well-known mathematical formulas, but the nomograph of **fig. 1** will provide the answer much quicker. Here's how it's done.

In **fig. 1A**, a line is drawn between the two known values: 5 mH and 1 kHz. This is labeled 1 in **fig. 1A**. The intercept of line 1 on the $X_L$ scale of **fig. 1A** shows the inductive reactance of this combination to be somewhere between 10 and 100 ohms.

Moving to **fig. 1B**, a line is drawn between 5 on the L scale and 1 at the top of the F scale. This location of line 2 was

*Used by permission of C.M. Marcott, editor, *Electronic Products Magazine,* United Technical Publications, New York.

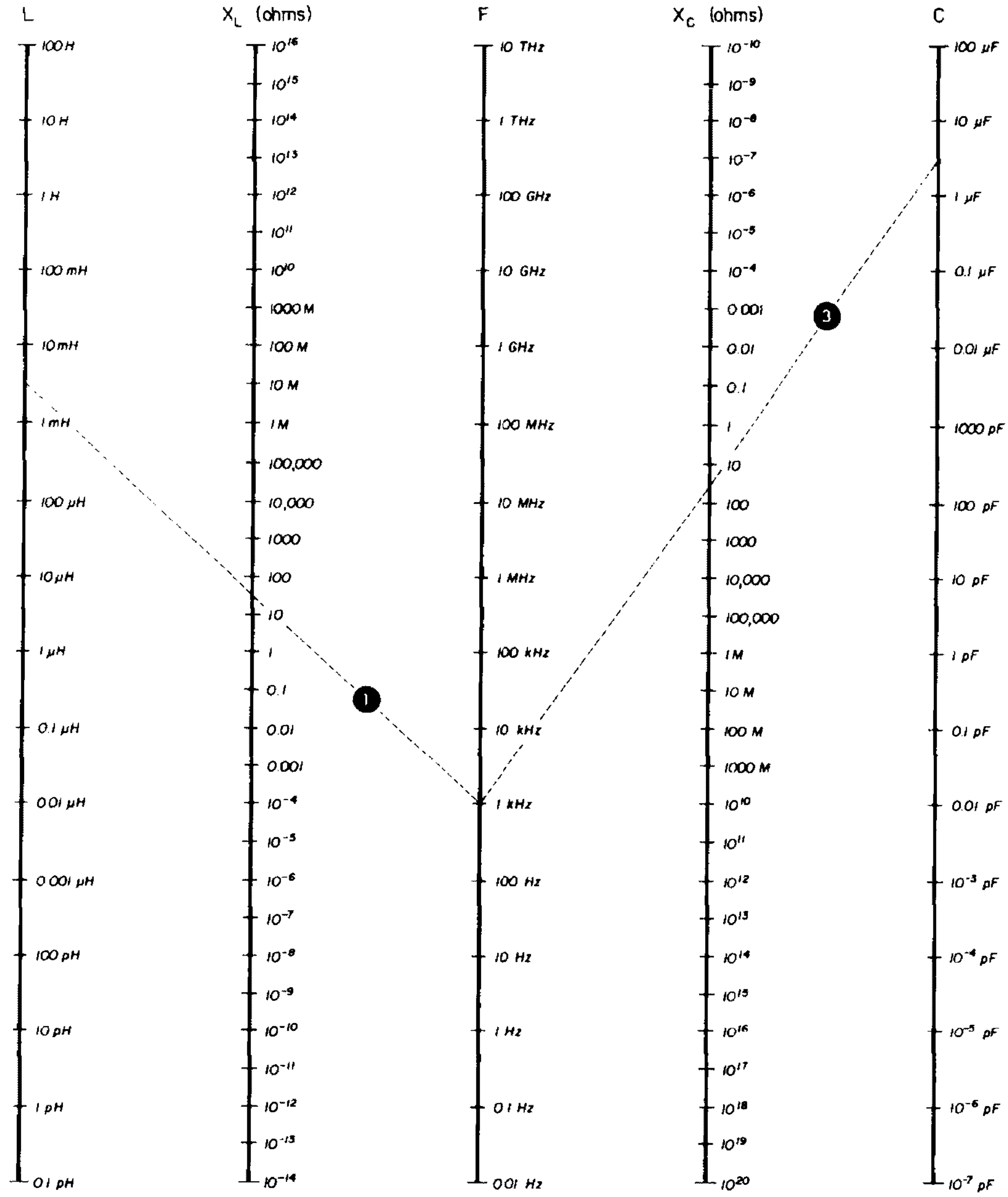



**fig. 1A. Reactance nomograph. This chart is used to determine magnitude and decimal location; significant figures are found from fig. 1B.**

purely arbitrary; the line, in this case, could just as easily have been drawn between 5 on the L scale and the bottom of 1 on the F scale.

The intercept of line 2 on the $X_L$ scale of **fig. 1B** is what's important. It intercepts the $X_L$ scale at approximately 3.2. This is the significant figure, or decimal multiplier, for the value determined from **fig. 1A**. The inductive reactance is therefore 10 (from **fig. 1A**) multiplied by 3.2, or 32 ohms.

## determining capacitance

Returning to **fig. 1A**, a line is shown between 1 kHz on the F scale and 32 ohms on the $X_C$ scale. This is labeled 3 in **fig. 1A**. Line 3 intercepts the C scale at 5 μF, which is the desired capacitance for C1 of **fig. 2**.

An inductive reactance of 32 ohms is used to find C1's capacitance, because basic theory says that resonance in a tuned circuit requires that $X_C$ equal $X_L$. This is the principle behind the calculations shown here. The nomograph can be used to solve other reactance problems as well.

## useful hints

When working with nomographs, the accuracy of the final result will depend

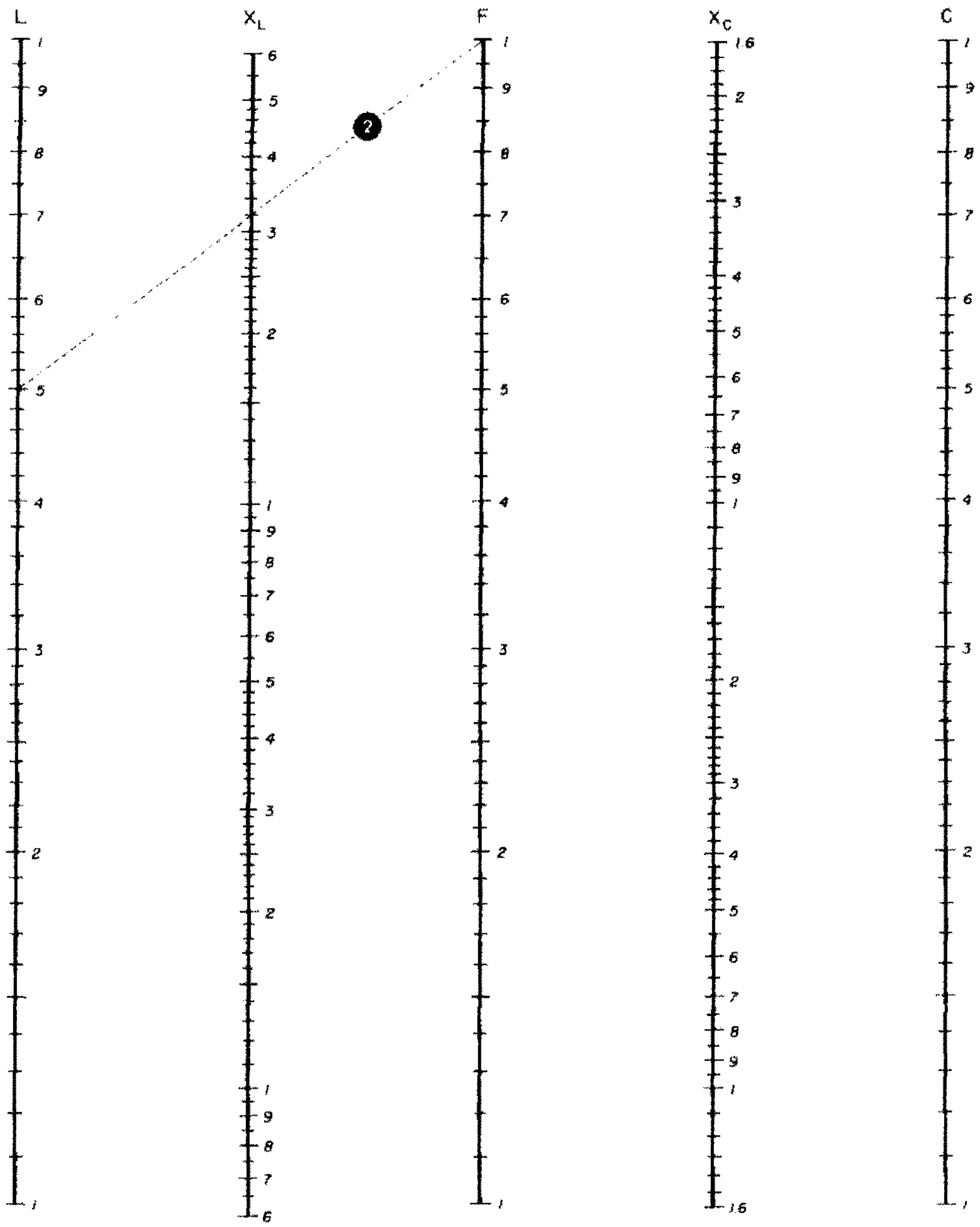



**fig. 1B. Reactance nomograph. This chart is used to find significant figures after magnitude and decimal location have been determined from fig. 1A.**

on how accurately you draw the connecting lines between the unknowns. Many nomographs give you a "ballpark" answer. If you wish to refine the result, you'll have to use mathematics. If used with care, a nomograph is a great time saver and can provide answers with accuracy sufficient for most practical problems.

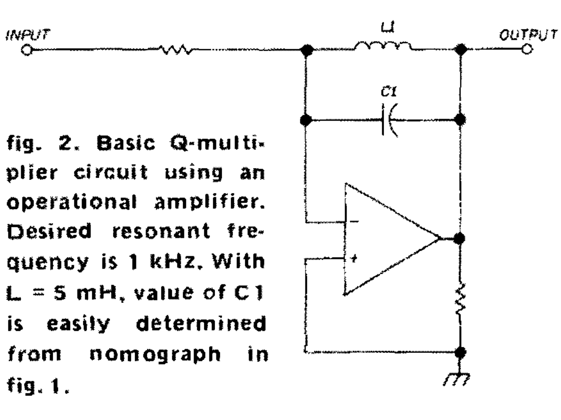



**fig. 2. Basic Q-multiplier circuit using an operational amplifier. Desired resonant frequency is 1 kHz. With L = 5 mH, value of C1 is easily determined from nomograph in fig. 1.**

A sharp pencil or draftsman's dividers should be used to locate the end points of the two unknown variables. Lay a straight edge against one leg of the dividers, and rotate the straight edge until you pick up the other point. Read the value at the intercept and mark it down on a piece of scratch paper. This will avoid cluttering the nomograph, which can then be used indefinitely as a computational tool.

**ham radio**

# parasitic oscillations in high-power transistor rf amplifiers

Robert C. Wilson, WØKGI, 6577 S. Newland Circle, Littleton, Colorado 80120

More than one design engineer has gained a few grey hairs trying to clean up the output of his 50- or 100-watt transistor transmitter. The reason is parasitic output frequencies, which were not mentioned by the textbooks and which may have been included in the output-power rating by the transistor manufacturer.

The subject of parasitic output has been avoided whenever possible by device salesmen, but in reality it's the one big reason why transistor transmitters aren't found in great profusion. Transistor parasitics are unlike tube oscillations, and the best solution to the problem is still to be found.

## transistor parasitic oscillations

There are three types of transistor parasitics. The first is the free-running type. This may be due to tuned or semituned circuits that self-oscillate at frequencies unrelated to the amplifier driving frequencies. The solution is the same as in tube design; that is, the problem circuit is reduced in Q or detuned. The second and third types of parasitics are much more insidious and difficult to eliminate.

The second type is due to the parametrically pumped characteristics of the transistor and is produced as submultiple frequencies of the amplifier driving frequency. These are very difficult to detect, as they are exactly locked to the drive frequency and are normally outside the range usually checked for spurious response.

The third type is low-frequency noise amplification, due again to parametric pumping. The low-frequency noise is not harmonically related to the amplifier drive frequency, but may be of either the so-called 1/f origin or from any low-frequency circuit that can be pumped into oscillation. This type of parasitic may be noticed as a rough-sounding spurious signal in the vicinity of the desired signal, or as white noise centered symmetrically around the desired frequency. If type 3 occurs it is low in frequency, but it will modulate the desired signal in the same way as any modulator, except with highly undesirable results.

The last two types may be detected indirectly by sudden increases in power output, or as changes in collector current while tuning. Because most people tune for maximum output as indicated by a power meter, most likely these spurious frequencies will be maximized.

## stabilization methods

Despite the fact that it has often been advocated, detuning the amplifier is no

solution to the problem. The real solution is to prevent the occurrence of the parasitic under any condition. To date this has not been possible, but several techniques will make rf amplifiers reason-

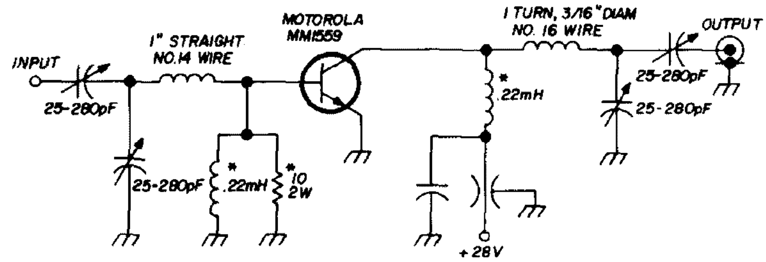



fig. 1. This is the brute-force method of stabilization, where pumped frequencies are swamped by shorting out low-frequency power. Values shown are for 150 MHz. Components marked with an asterisk are used to stabilize the amplifier.

ably stable. One method, recommended widely, is to prevent low-frequency response in the amplifier by using very small values of inductance in the base return and collector dc feed. **Fig. 1** is an example. A problem occurs when the frequency is reduced in that the chokes must be larger, and often the expected result happens; that is, parasitics.

Tube-type neutralization networks are generally ineffective. The reason is that

| | |
|---|---|
| L1 | 10 turns no. 18 on 1½" toroidal core, Indiana General type Q-1 |
| L2 | 3 turns no. 18 on 2" toroidal core, Indiana General type Q-1 |
| T1 | Primary 7 turns no. 18, secondary 3 turns no. 18 on 1½" toroidal core, Indiana General type Q-1 |

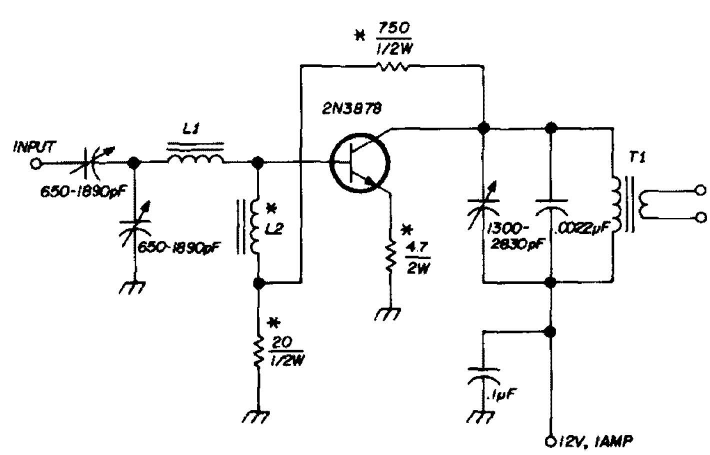



fig. 2. A 160-meter stabilized amplifier using various methods of reducing types 2 and 3 pumped parasitics. Components marked with an asterisk are used to stabilize the amplifier.

the transistor capacitance varies at an rf rate, while the neutralization components are constant. A method I've found useful is to suppress low-frequency gain by resistive feedback **(fig. 2)**. The cost is a small amount of lost rf gain (perhaps 1 dB) and a somewhat increased dc-circuit current. The thing to remember is not to bias the transistor full-on in the quest for feedback. The result in a noncurrent-limited circuit is the immediate burnout of the transistor—a very costly fuse!

Summarizing, at least three types of rf transistor parasitics occur. These often go unnoticed, because they're not at expected frequencies and are sometimes produced in ways foreign to tube engineers. So far, techniques for parasitic reduction leave much to be desired as they are brute-force methods rather than elegant solutions to the problem. Clean, high-power rf amplifiers are a possible but ticklish proposition and await some yet-undiscovered technique to produce optimum results.

## bibliography

1. R. Minton, "Design Trade-offs for RF Power Amplifiers," RCA Publication ST-3250, Sommerville, N. J., 1966.
2. "RCA Silicon Power Circuits Manual," RCA, Harrison, N. J., 1967.
3. J. E. Tatum, "VHF/UHF Power Transistor Amplifier Design," Application Note AN-1-3, ITT, West Palm Beach, Fla.

ham radio

# cw transceiver operation with transmit-receive offset

It's easier to operate cw with a slight frequency offset between direct-conversion transceivers— here's why

Richard S. Taylor, W1DAX, 54 Slade Street, Belmont, Massachusetts 02178

Recently while playing with a design for a direct-conversion cw transceiver, I had some thoughts about the significance of transmitting and receiving on slightly different frequencies, as this design requires. No one seems to have given this idea much thought, although many apparently operate this way all the time. My first reaction was that there might be some situations in which the offset would cause real problems—for example, two identical transceivers trying to talk and getting nowhere because one of them was tuned to the wrong side of zero beat.

My conclusion is just the opposite. The offset adds no new problems to the operation of the transceiver. In fact, if the transceivers are identical (and the offsets the same) it's easier to operate with the offset than without it.

## frequency offset

First, the frequency offset occurs in a direct-conversion transceiver (**fig. 1**) because the vfo must be offset from the received signal frequency to produce an audio beat note. Since the transmitted signal is just the vfo output amplified, the transmit and receive signals are different by the frequency of the beat note. **Fig. 2** shows a response curve of a cw direct conversion transceiver (dct) and how two other dcts might communicate with it.

In **fig. 2**, the rf response curve has the form of the audio filter plus its mirror image. The mirror image occurs because signals are audible on either side of zero beat.

Stations anywhere under the curve can be heard. In looking for a contact transceiver one, say, tunes for a signal above zero beat and responds. The home rig will hear him at $f_2$. Transceiver two tunes for a signal below zero beat, home hears him at $f_3$. Communication is possible in either case. Note that the beat frequency, although different in the two cases, is the same between pairs of stations. When home and station one talk, they both will have a beat frequency of $f_{b1}$; when home and two talk, it will be $f_{b2}$. This requirement for a common beat frequency may be disconcerting to operators whose ears peak at different frequencies, but it doesn't prevent contact from being made.

Operating with another station that's not a transceiver is just as simple and proceeds as above, except that now each station can choose a favorite beat note because of the bfo control at the non-transceiver station.

## calling CQ

In answering a CQ with a transceiver the home station does not zero beat, because he can't hear the calling station.

The home station must tune the calling station to one side or the other of

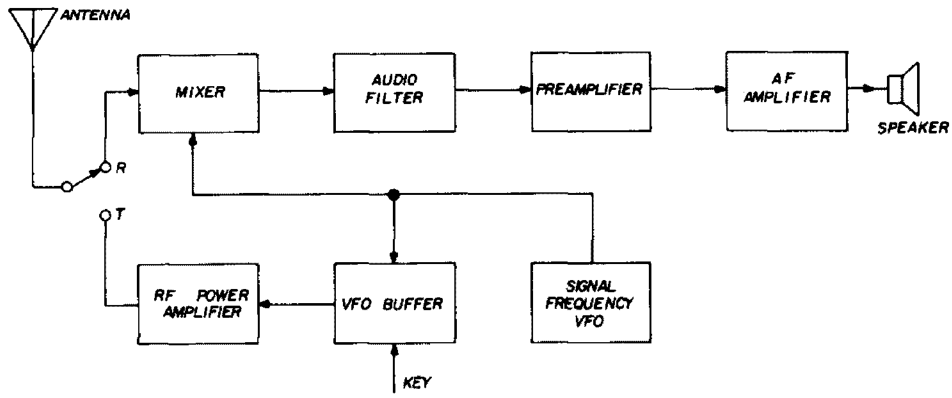



**fig. 1. A basic cw direct-conversion transceiver. The vfo is common to receiver and transmitter.**

zero beat. The choice will probably be made on the basis of least interference. If the other station is also a transceiver, communication will automatically be established. If the other station is not a transceiver, the offset probably won't be noticed as he tunes for replies.

In calling CQ with a transceiver, however, the replying station, if not a transceiver, will probably zero beat the transceiver carrier. The transceiver would then have to be retuned to hear the reply. On the second go-round the replying station might think the transceiver has drifted (it has, of course), but as long as the replying operator retunes only his receiver and leaves his vfo alone, communication will be established on the second try, and no further retuning would be necessary.

If the transceiver were retuned to reply to a crystal-controlled station, no contact would be made. This is, of course, the same thing that would happen with an ssb transceiver. The solution is to listen around your own frequency and wait for people to come to you, as is done on ssb.

So the offset need not be a problem. At least it should be no more of a problem than the inability to listen on any frequency but one's own, which is the characteristic of any transceiver.

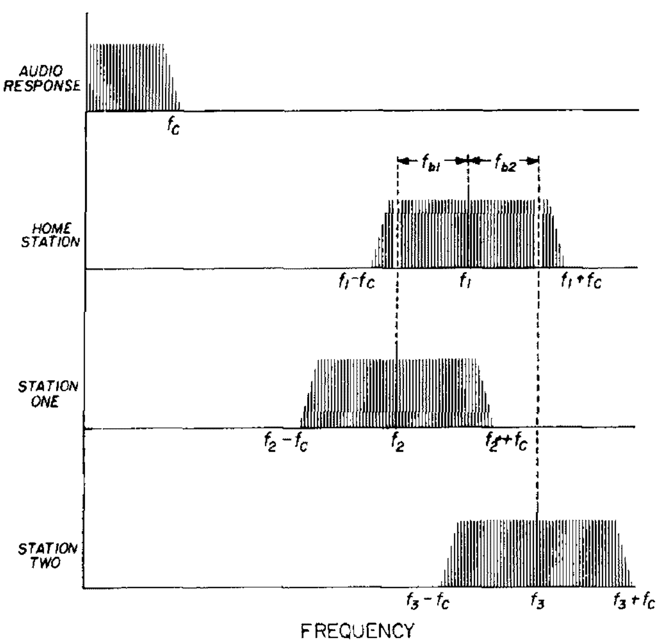



**fig. 2. Response curves for three dct's. The audio response for each rig is shown on the first line. The beat frequencies are the difference between carrier frequencies in each case: fb1 = f1 — f2 and fb2 = f3 — f1.**

## what about audio selectivity?

Surely, the wide-open bandpass shown in **fig. 2** is unsuitable in today's conditions. Audio peaking can improve receive selectivity, but this introduces a problem **(fig. 3)**. Here, both the home station and station one have added audio selectivity, but at different frequencies. The home station is peaked at $f_{b2}$; station one is peaked at $f_{b1}$. Let's say station one is calling CQ and the home station hears him, tunes him to his audio peak, and replies. But home station's transmitting frequency is outside station one's audio peak and, therefore, may not be heard. Avoiding this situation requires either opening the bandwidth of both transceivers, or standardizing the offset between them.

*I recommend offset standardization by means of identical peaked audio filters at*

*the two transceivers.* Not only is improved selectivity achieved, but a system advantage is gained as well. Tuning is simpler. Once home tunes the incoming signal to his audio peak, communication is optimized, because the signal transmitted from the home station will be at the audio peak of the other transceiver. One envisions highly selective transceivers with but one frequency control, which is attractive.

## bandwidth considerations

The best shape for the peaked bandpass requires careful consideration. A sharp peak is desirable from the standpoint of both offset standardization and selectivity. It would be undesirable to have too accented a peak, however, as it would be hard to hear stations with offsets different from your own. Possibly 6–8 dB would be a good number for the peak above the low-frequency response level (**fig. 4**). An optimum width might be a few-hundred Hz at the 6-dB points. An ability to hear stations below the peak (nearer zero beat) is desirable, so that stations which tried to zero beat your carrier would be audible. This would make the direct-conversion transceiver a good performer when working all kinds of rigs—not just when working other transceivers.

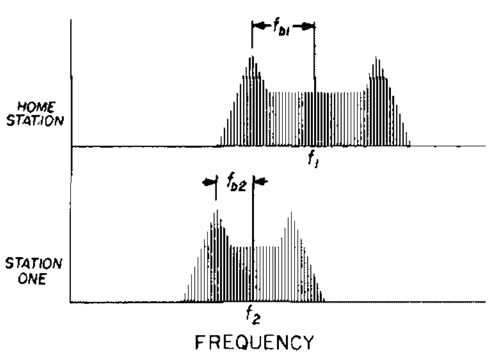



**fig. 3. Two dct's with audio selectivity at different frequencies. Note that when the home station centers his audio peak on station one's carrier at f2, station one's audio peak is way off the carrier at f1. In this example station one wouldn't hear the home station at all.**

It's entirely possible to build a direct-conversion transceiver using phasing techniques that would provide true single-signal reception with a response on only one side of zero beat. It would look much like the direct-conversion ssb receiver I described in reference 1. I don't believe there is any advantage to doing this, however, because the sideband that the single-signal receiver ignores may well be the one on which the responding station replies. If such transceivers became common, it would be necessary to designate which sideband was to be used on each cw band to ensure that the transceivers would communicate. The dsb version; i.e., with response on both sides of zero beat, seems the best choice, at least at present.

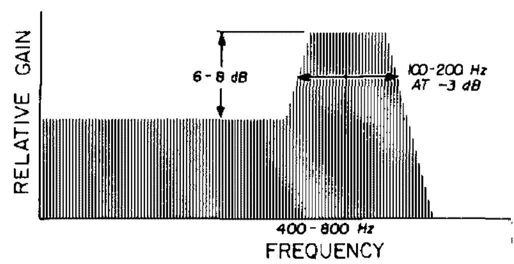



**fig. 4. A possible standard passband for cw dct's. The optimum numbers would be determined by in-use testing.**

## conclusions

**1.** Transmit-receive frequency offset is not a hindrance to cw communication, whether between two transceivers or a transceiver and regular stations.

**2.** Frequency offset standardization is required if good cw selectivity is to be obtained.

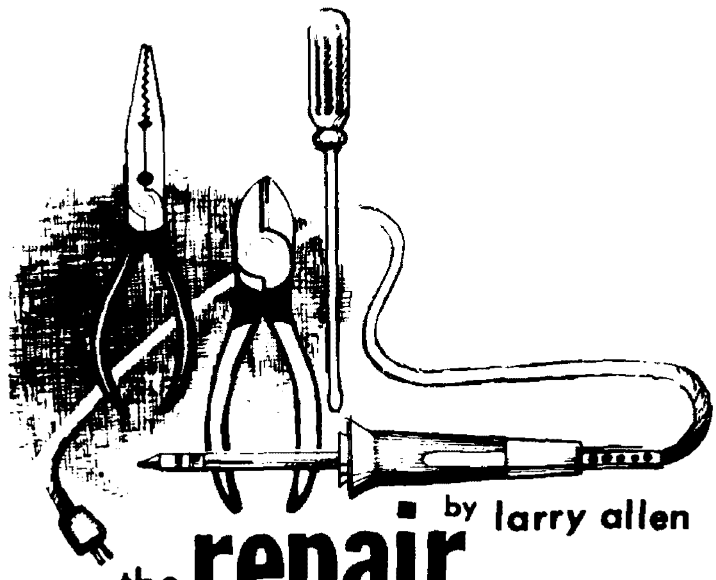
**3.** This standardization is best achieved by means of a standard peaked filter response to be used in all cw transceivers.

Maybe the cw dct would be the thing to get a lot of us back on the cw bands. Certainly the simplicity is attractive.

### reference


1. Richard S. Taylor, W1DAX, "A Direct-Conversion S.S.B. Receiver," *QST*, September, 1969, pp. 11-14.


**ham radio**



■ by larry allen

# the repair bench

## finding faults in rf and i-f amplifiers

**What do you do if you** can only pick up nearby or strong stations? What if they come in weak and perhaps noisy? First thing you should do is suspect the rf amplifiers in your receiver.

This isn't unusual, particularly with transistor front ends. If the manufacturer (or you, if the receiver is homebrew) designed the rf amp to use the sensitive but delicate field-effect transistor, a gate punctured by static surge isn't at all uncommon.

The problem lies in recognizing a bad front end. If the mixer and i-f stages are naturally quiet, and transistor stages often are, you may not know an i-f failure from an rf one. Or the fault may be in the automatic gain control (agc) system. Only careful testing will tell you for sure.

### amplification vs noise

Familiarity with your receiver is the best assurance of knowing when there really is trouble. You should get to know how much natural receiver noise to expect when there's no station.

Then, when suddenly you can't pick up stations you know should be there, make a listening test. Tune the receiver dial to an empty spot. Turn rf and af gain up.

Is receiver noise (the background thermal hiss) up to snuff? If so, the i-f stage must be amplifying. Also, the mixer stage is probably okay; much front-end thermal noise normally originates there.

Yet, some of today's field-effect transistor (fet) front ends are too quiet for this kind of analysis. It's normal to hear almost no receiver hiss. You have no choice then but to rely on other testing methods. You try to determine rf-stage sensitivity. Or, you can just test the stage by regular dc-measurement methods.

### what's in a rf stage

Most of what I say about troubleshooting rf stages can be applied to i-f stages, too. There's little difference.

In most a-m receivers, rf stages tune over several different bands. The i-f stages are fixed-tuned. But it's common to tune ssb receivers nowadays by synthesis—the same as the transmitter. In that case, the receiver rf stages are fixed-tuned. You troubleshoot them the same as i-f stages.

A typical old-fashioned tube-type rf stage is drawn in **fig. 1**. Only one deck of the bandswitch is shown; the input band coils are omitted for simplicity.

Most hams can figure out a way to track down trouble if they are sure what a stage is supposed to do. I don't mean in just a general way, but specifically—each part of the stage. Take the stage in **fig. 1** for an example.

First and foremost are the *input* and *output* circuits. T1 and C2 are the input

coupling components. R1 is the input load, decoupled by C3. Ordinary signal tracing or injection is the way to make sure these components work.

The output circuit is T2, decoupled by C7. Those decoupling components are an important part of the input and output circuits, so don't forget them when you analyze stage operation.

There are also tuned circuits to contend with, although part of them are not lyze the screen or cathode dc circuit.

The cathode dc circuit, which can be considered dc supply even though it's just a ground return, is through R2 and R3. C4 or C5 can become part of this circuit if either happens to short. Otherwise, they don't affect cathode dc voltage.

Notice that R3 is variable. It's the *rf gain* control for the receiver. (In the receiver from which this example is taken, R3 also is part of the i-f stages.

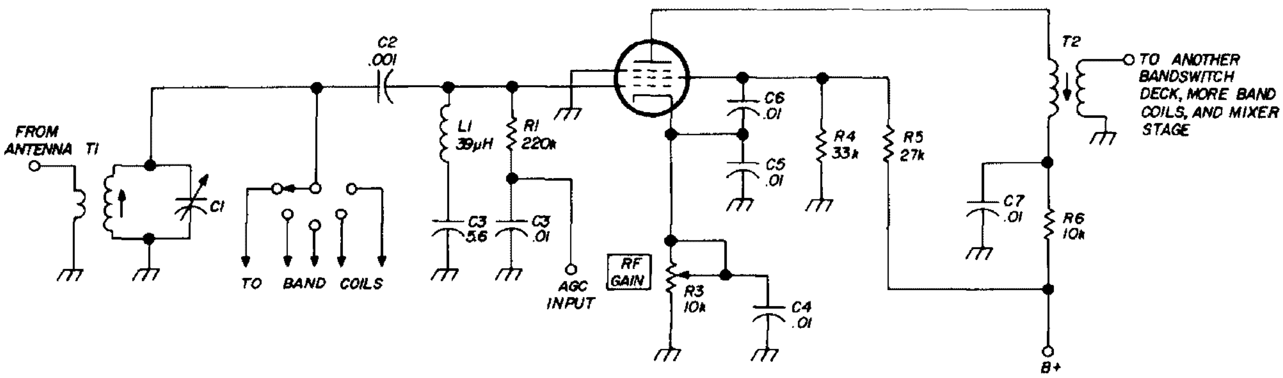



**fig. 1. Rf stage contains circuits that must be checked individually if you don't use test procedures that check overall stage performance.**

shown in **fig. 1.** T1 and C1 make a tuned circuit. But they are in parallel with band coils you can't see. The band coils, with the inductance of T1, set the band the stage is to tune across and C1 tunes the specific frequency. If there's a fault in one of these coils, that band won't tune properly. If the fault is in T1 or C1, *none* of the bands tune as they should.

Next, consider the *supply* circuits. They carry dc voltage to the tube elements.

The plate supply is through R6 and the primary of T2. Capacitor C7 is important in the plate dc supply circuit only because of the possibility it might short. You therefore must consider it part of the plate supply circuit when you're diagnosing.

Resistors R4 and R5 are the chief components of the screen supply circuit. C6 is a potential part of it—if the capacitor happens to become leaky or shorted. Leakage in C6 would put voltage intended for the screen onto the cathode. Consider that possibility when you ana-

That connection is omitted here for simplicity.) Changing the value of R3 between R2 and ground varies cathode bias on the tube. The pentode is a sharp-cutoff type; its gain depends sharply on its bias. Thus, by changing bias, R3 controls rf amplification.

Voltage at the grid is controlled from the agc stage. The agc control voltage is fed through R1. Decoupling capacitor C3 is part of the grid-supply circuit only if it shorts or gets leaky. C2 and C3 might become part of that circuit if either went bad.

There's another resonant circuit. It isn't tunable. It may also be called a trap circuit, because that's what it's there for. L1 and C3 form it, and it's resonant to 9 MHz, the i-f of this receiver. It traps out any stray 9-MHz signals, preventing them from being amplified and running through the mixer to upset the i-f stages.

That about sums up the circuits in this rf stage. Input, output, tuning, bypass or decoupling, and trap: those are the signal-carrying circuits. Plate supply, screen

supply, cathode return, and grid bias: those are the dc-carrying circuits. You have to consider each when you set about troubleshooting an rf stage like this one.

## transistor rf stages

The transistor input stage from one receiver is drawn in **fig. 2**. You can probably identify the circuits. They may look a bit different from those in fig. **1** because they're in a transistor stage.

The input circuit comprises L1, C1, C2, and both R1 and R2 as load resistors. Decoupling for the two resistors isn't in the diagram, but it's understood. The agc line has a bypass capacitor not shown; so does the 12-volt supply.

L1 and C1 are fixed-tuned, although adjustable with a tuning tool. They form a broadband tuned circuit. (This particular rf stage is part of a one-band transceiver.) The taps on L1 are for impedance matching.

The output circuit comprises L2, C4, R5, and C5. L2-C4 are a tuned tank, with R5 as a band-broadening load across it. L2 is adjustable for band-peaking. C5 couples amplified rf energy to the mixer stage (through another tuned circuit, omitted for simplicity).

The only other signal circuit is emitter bypass capacitor C3. If it opens, substantial degeneration can occur, but it only makes the stage weak—it doesn't make it dead.

There are only three dc supply circuits. That's because a transistor has only three elements to receive voltage.

The transistor is pnp. Therefore, normal forward-bias operation puts the emitter positive, the base less positive (same as more negative than emitter), and the collector far less positive (same as far negative from the emitter).

The emitter gets voltage directly from a positive 12-volt supply line through R4. C3 is the decoupling capacitor, and is a concern to the dc circuit only if it shorts or gets leaky.

Collector goes to ground through L2. The winding has no appreciable dc resistance, so for dc the collector is grounded. That puts it far negative with respect to emitter.

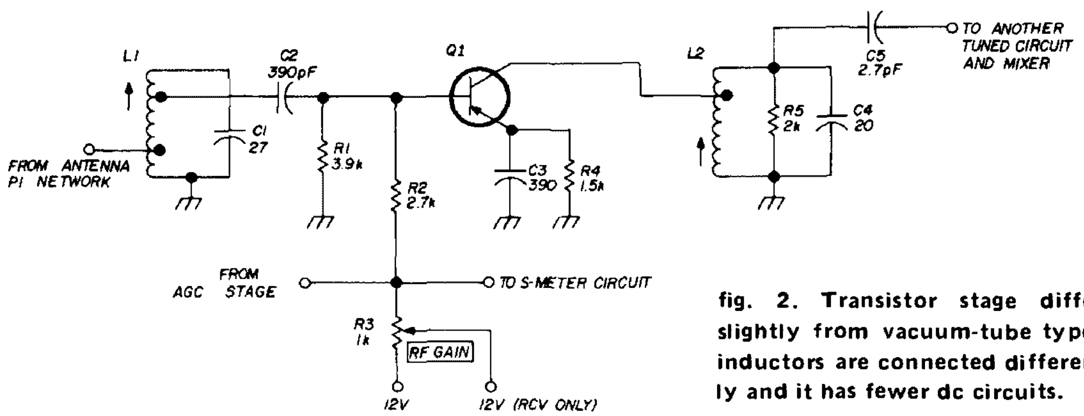



**fig. 2. Transistor stage differs slightly from vacuum-tube type—inductors are connected differently and it has fewer dc circuits.**

The base has the only complicated supply network. The main dc comes from the 12-volt line through R1. However, a connection through R2 lets the actual voltage—and therefore bias on the transistor—be varied by the agc line and by the setting of *rf gain* control R3. The transistor operating characteristic is such that bias controls amplification. Thus the *rf gain* control sets optimum gain of the stage, and agc varies it to accommodate signal strength.

Of course, C2 is part of the base dc circuit only if it comes defective. A faulty decoupling capacitor in the agc line could also affect base bias. You need remember these capacitors only if you're troubleshooting and find the base voltage is wrong.

## external influences

One other thing you can't forget when

you're troubleshooting rf and i-f stages. A trouble in the stage may be caused somewhere else.

A blocked i-f or rf stage is common. Sometimes, that's traced to an agc stage overdoing its bias thing. Or, the i-f or rf amp may block on strong signals—a sign of overload. That, too, may be traceable to a faulty agc stage—this time not doing enough.

The receiver block diagram is sometimes helpful in spotting stages that affect rf or i-f. Some don't show that much detail and you have to rely on your ability to read the schematic. **Fig. 3** is a partial block diagram of the set from which **fig. 2** is taken. Its detail is enough to be helpful.

You might find the voltage upset in an i-f stage, yet the agc stage works normally. Suspect the s-meter hookup. If any part of that circuit shorts to ground, it could foul up bias on rf or i-f stages. A short inside the s-meter might make the *rf gain* control work wrong.

In other words, examine the schematic or block diagram before you go tearing into any rf or i-f stage. If external stages affect the rf stage, check them out or isolate them from the rf stage some way.

## testing rf amplification

One way to see if an i-f or rf stage is working right is to measure its gain. In modern transistor receivers you can expect to find a voltage-gain factor of 20 or more. A tube stage usually gives even higher gain.

You can make this measurement fairly easily if the output of your rf generator is calibrated. First, clamp the agc line with whatever dc voltage produces normal idling (no-signal) bias on the rf or i-f stage you're testing. That bias is usually written on the schematic or on the voltage chart.

Clip your vtvm to the a-m detector output, or through an rf probe to the output of the last i-f amp. Feed the generator signal to the *input* of the rf or i-f stage being measured. Note the dc meter reading. Set the generator output level for some meter reading that's easy to remember. Make a note of the rf output level of the generator.

Move the generator signal to the output of the stage being tested. Turn up the generator level until the meter reads the same as before. Divide the new generator output-level reading by the earlier one. The result is the voltage gain of the stage.

As an example, suppose 0.7 microvolt of signal drives the meter to 1 volt dc when the generator is connected to the stage input. When you connect the generator to the stage output, you have to turn the generator up to 14 microvolts to get

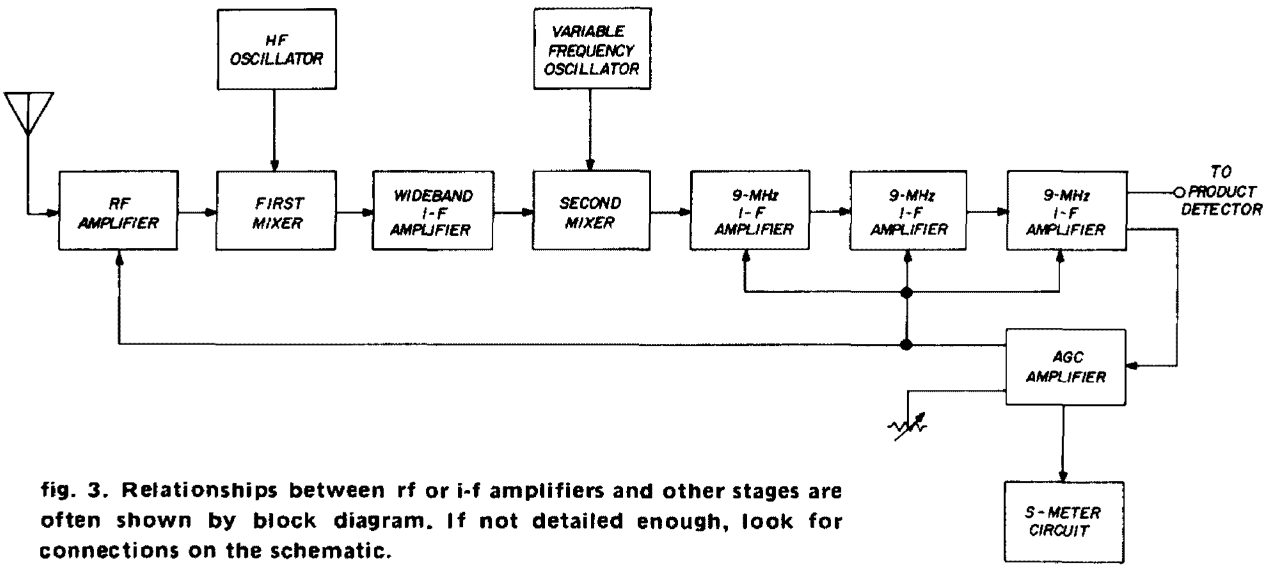



**fig. 3. Relationships between rf or i-f amplifiers and other stages are often shown by block diagram. If not detailed enough, look for connections on the schematic.**

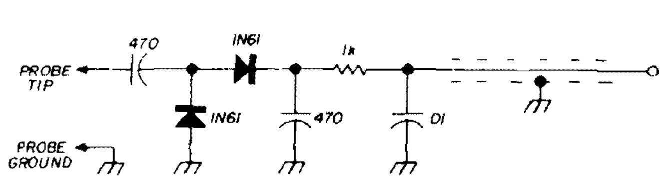
that 1-volt dc reading on the meter. Dividing 14 by 0.7 gives 20. That's the voltage gain of the stage.

Unfortunately, only the more costly signal generators have calibrated output. You may have to use a less accurate way. It's only relative, and your best bet for using it is to make measurements while your receiver is working normally and record them for reference.

You'll need a doubler-type rf probe for your vtvm. A suitable circuit is sketched in **fig. 4**. It's more sensitive than the ordinary single-diode probe. The vtvm should be a very sensitive one, with lowest full-scale reading 1.5 volts or less.

Again, clamp the agc. Turn the *rf gain* control wide open. Keep the generator unmodulated.

**fig. 4. Circuit for a voltage-doubling probe you can use with a vtvm to measure relative rf signal levels in rf and i-f stages. Entire probe should be shielded to prevent hand capacitance from upsetting the reading.**




Feed the rf signal to the antenna input jack. Tune the generator to the center of the band if the stage is fixed-tuned; if the stage is tunable to one frequency, set the generator precisely to that frequency.

Connect the vtvm probe first to the base of the rf transistor. Set the meter on its lowest range. Turn up the generator signal until you get a perceptible reading on the meter. Then set the generator output for some very small but definite voltage indication—say 0.01 volt. Don't change the generator setting.

Move the vtvm probe to the base of the mixer. The reading should be much higher now—say nearly 0.2 volt dc. That represents a gain of about 20 if you're using the doubler probe.

Obviously, these figures are approximate. Voltage gain for a tube is roughly the same. For tubes or transistors, however, the surest system is to make a record of normal amplification while your receiver is new. Then use the same measurement method when you test on the repair bench.

## dc troubleshooting

You should already know the methods of tracking down the cause of incorrect dc voltages on an rf stage. You might want to make dc tests before you go to the trouble of putting the rf probe on your vtvm.

But usually you'll find that kind of fault is obvious—as when the stage is completely dead. It's for subtle weakness or abnormal overloading you need to clamp the agc and test stage gain. Then you can hunt down the small voltage problem—or bad transistor or tube—that's causing the trouble.

## checking the stage another way

You can also use a form of signal injection. Connect the vtvm, without a probe, to the a-m detector of the receiver. If it's handier, keep the probe on the vtvm and connect it to the i-f input of the product detector (or at the output of the last i-f amp). Clamp the agc as before.

Connect the rf signal generator, tuned to the rf frequency as already described, to the input of the mixer stage. Turn up the generator output just enough to cause a reading on the meter. Make a note of the reading.

Move the generator back to the input of the rf stage. Note the increase in the reading. The dc-voltage increase should be about the same as the one already described. Without the doubler probe, a 10-times increase means about 20 gain in the rf stage. With a doubler probe instead of the set's a-m detector, a 20-times increase means roughly 20 gain in the rf stage.

Now go a step further. Be sure you've got the generator frequency set precisely. Try tuning the coils in the tuned circuits. If the meter reading doesn't vary, the coils may be at fault. Before you replace

them, though, make sure the capacitor that decouples each coil is not open.

Finally, if tuning is erratic and you can't seem to make head nor tails of how the coils tune, check the bypass capacitors on the supply lines and on the agc line. One of them may be open.

## looking to the future

Remember that the ways of checking rf stages outlined here can be used just as well with i-f stages.

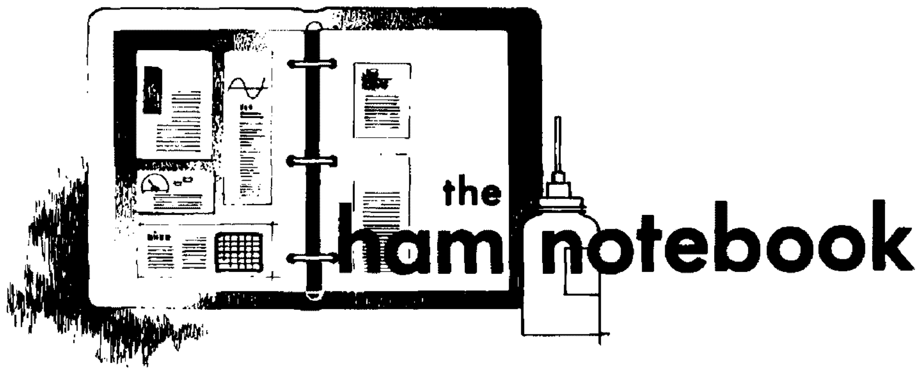
There's still another way of troubleshooting rf and i-f stages: with sweep alignment. Unfortunately, no inexpensive sweep generator available today goes down far enough in frequency.

In a future column I'm going to show you how to make your regular sweep generator go down far enough to sweep 60-kHz, 455-kHz, and other i-f amps. You can do it without modifying the instrument you have.

First, though, there's a new troubleshooting system that has come to my attention. It's called *1-2-3-4 Servicing* by its originator, Forest H. Belt. In the next issue of repair bench, I'll tell you what *1-2-3-4 Servicing* is all about. That'll prepare you to understand what goes on when you use the sweep-alignment method of rf and i-f troubleshooting.

ham radio


"Don't ever let it slip to my wife that field day isn't the second and fourth Saturday of every month . . ."




## protection for solid-state power supplies

The current crop of Heathkit solid-state power supplies is great. I have three of them, which I use in ham work and in areas far removed from hamming. Some problems developed with these supplies, and the purpose of this note is to recommend a slight design change that you can make to protect your power supply. Even if your supply isn't a Heathkit—home brew, for example, but using similar circuits—you might consider this inexpensive way to add protection against external influences.

Many experimenters have surplus relays, especially the 28-volt type with coils that require about 200 mA for operation. It is convenient to test them with a solid-state power supply. The stored energy in the relay coil can develop quite a wallop when the circuit is opened to de-energize the coil. Standard practice is to connect a diode across the coil to suppress the stored voltage. However, when checking many assorted relays, the tendency is to go the easy way and take a chance that the switch gap, upon opening, will dissipate the stored energy. You might get away with this a few times, but not for long. Sooner or later a large negative voltage spike will back up into the power supply, and you've got problems.

The easiest way to protect the supply in this case is to add a diode across its output terminals. Solder the diode to the lugs on the back of the binding posts, cathode to positive; anode to negative. A 1N4003, 1N538, 1N645 or similar type will work. Any negative transients trying to sneak back into the supply will be shorted to ground at the terminals. So much for protection from one external influence.

Another external influence is a short circuit, often a dead short, across the supply. Heath has a dandy circuit that cuts off the current when a short circuit is sensed. Upon removal of the short, current is restored, and no harm is done. The Heath circuit contains two supplies, each mutually independent. One is the heavy-duty supply, which provides power to the load. The other is a zener-regulated, constant-voltage reference source. Across this source is a potentiometer that provides a variable voltage. The positive terminal of the reference source is tied to the positive terminal of the power supply. The pot output is applied to the base of an error detector/amplifier transistor. During normal operation, the unregulated supply will attempt to match the level of the reference-supply voltage. The error detector senses the difference between these voltages. The result is a slightly positive voltage at the error-detector transistor base. As the pot is varied, the output voltage will tend to follow the reference voltage.

Suppose a short circuit develops. The current-limiting circuit will reduce output current to near zero, reducing the output voltage to near zero. The voltage difference at the error-detector transistor base

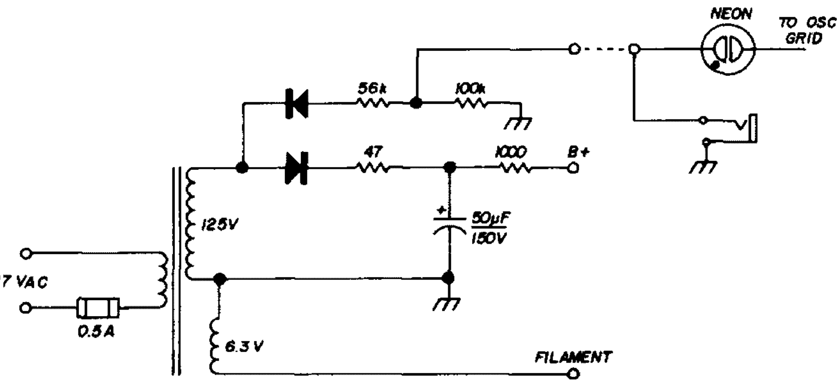



fig. 2. Isolation circuit for use with Heath HG-10B vfo and HD-10 keyer.

will be some voltage between the output voltage (zero) and the reference voltage—anything up to 35 V, for example. Since the reference voltage is negative with respect to common, the full negative reference voltage will appear at the error-detector transistor base, instantly zapping the transistor.

To prevent this disaster, Heath uses a diode in series with the error-detector transistor base. The diode will pass current in the positive direction only. However, if regulation is lost (this happened to me), the output voltage will shoot up to the full unregulated amount, say 50 V. As before, the error detector will sense the difference between output and reference voltages; except this time about +15 V will appear at the error-detector transistor base. Here the diode offers no protection; its job is to protect the supply in the event of a short circuit. The relatively high positive voltage on the error-detector transistor base will pass a high current through the base-emitter junction, and ZAP!

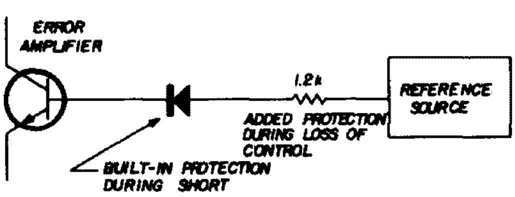



fig. 1. System used by Heath to protect their power supply. Resistor was added by author.

The answer to protection during loss of regulation is extremely simple. A resistor is placed in series with the error-detector transistor base, as in **fig. 1**, to limit base current. The cost in performance will be a slowdown in supply response—about 15 ns.

reference

1. Heathkit Manuals for Model IP-18 and IP-28.

**Frank Case, W3NK**

# independent keying of Heath HG-10B vfo

For owners of the Heathkit HG-10B vfo who might wish to make it independent of other equipment, the following idea will be helpful. My vfo is keyed with a Heathkit HD-10 keyer. The problem was how to use this model keyer with the vfo. The instruction manual for the keyer warns against using a voltage above −105 V.

The circuit of **fig. 2** shows how I solved the problem. All components except the transformer and fuse holder were mounted on a small piece of Vector board. The existing 4-conductor cable was removed and replaced with a length of ac cable. The fuse holder was mounted between the key jack and ac cable. The transformer, an inexpensive Japanese unit, was mounted vertically on the underside of the chassis next to the mode-switch wafer. The Vector board will fit nicely between the transformer and the rear chassis wall next to the terminal strip holding the neon lamp.

The connection from the key jack was removed from the cathode circuit and connected as shown. No switch was used, as I intended to leave the vfo on constantly.

**James H. Crouch, K4BRR**

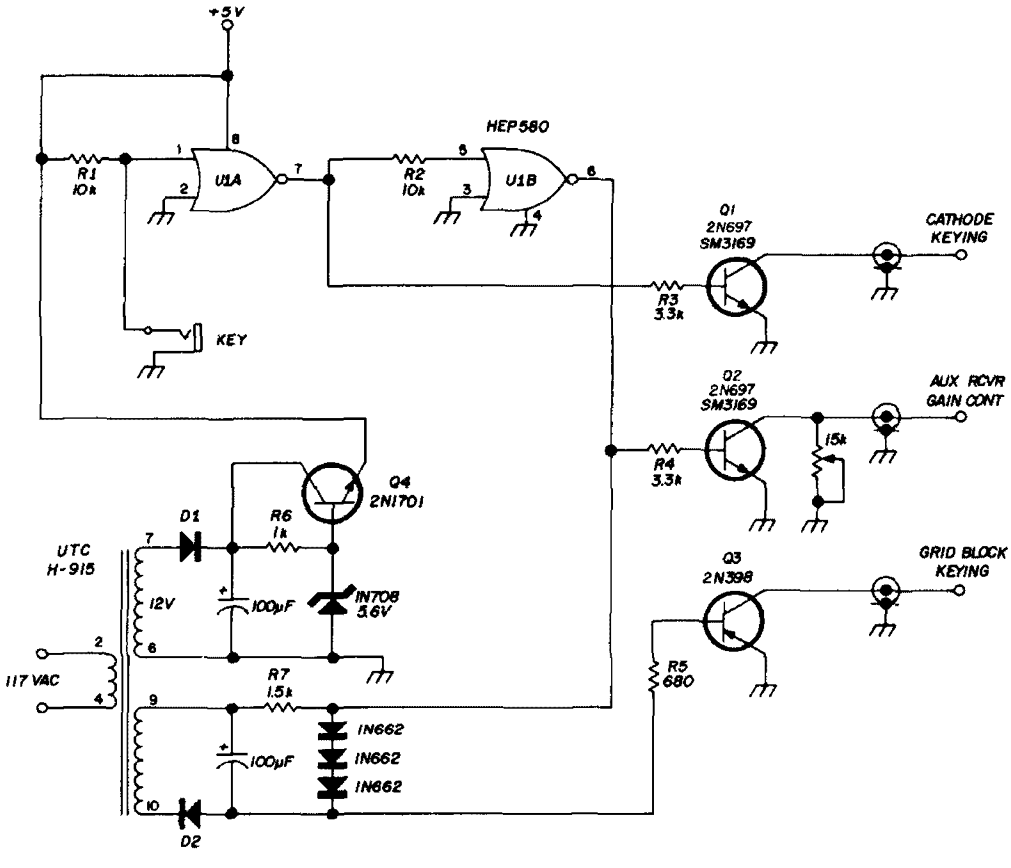



**fig. 3. CW break-in control circuit using ICs.**

## break-in control system

For really fast and effective cw break-in, many amateurs use separate transmitting and receiving antennas. With such an antenna system and fairly low power (under 100 watts output) the control circuit described here is all that's required for full and complete break-in operation. It uses the old idea of inserting additional resistance in series with the receiver gain control when the key is down to automatically decrease receiver gain and prevent overload. This method also allows the transmitted signal to be monitored in the receiver.

**Fig. 3** shows the circuit. When the key is up, U1A turns on, and the output at pin 7 goes low (0 volts). The output of U1B goes high because both its inputs are low. This turns Q1 off and Q2 on, so that the transmitter keyed circuit is open, and the auxiliary gain control is shorted to ground. The receiver now has full gain. Closing the key causes U1A to turn off, the output at pin 7 goes high, which forces the output at U1B pin 6 to a low state. As a result, the transmitter is keyed and the auxiliary gain control decreases receiver gain to its preset level.

Since I have two types of transmitters (cathode and block-grid keying), I added a negative source of voltage and Q3 to key the grid-blocked rig. With the key up, the positive voltage on U1B pin 6 overrides the negative supply at Q3's base, holding it off until the key is closed.

This new system replaces a dpdt relay and its keying circuit. In addition it provides for independent cathode and grid-blocking keying of different transmitters. I'm breaking only about 20 Vdc with Q1 and Q2, so the devices shown work fine. The 2N398 would be required for most grid-blocked keying systems.

The whole thing could be powered from a few dry cells, but the surplus transformer was available to make the system operate on 117 Vac.

**Cal Sondergoth, W9ZTK**

## ssb input source for vhf, uhf transverters

Many amateurs require a good vfo for use in the vhf/uhf bands. The 28-MHz output of an ssb transceiver or transmitter can be used as an input source for an up-converter to obtain the desired vhf/uhf frequency. It's highly desirable to maintain the original modulation system of the transceiver or transmitter.

For output on 144 MHz, one could use 166 + 28 MHz; for 432-MHz output, the conversion can be made by mixing 404 and 28 MHz. The best method to obtain output on 1296 is to use 28 + 518 MHz, then mix the 546-MHz resultant signal with 750 MHz. This method will avoid all frequency components except the desired one.

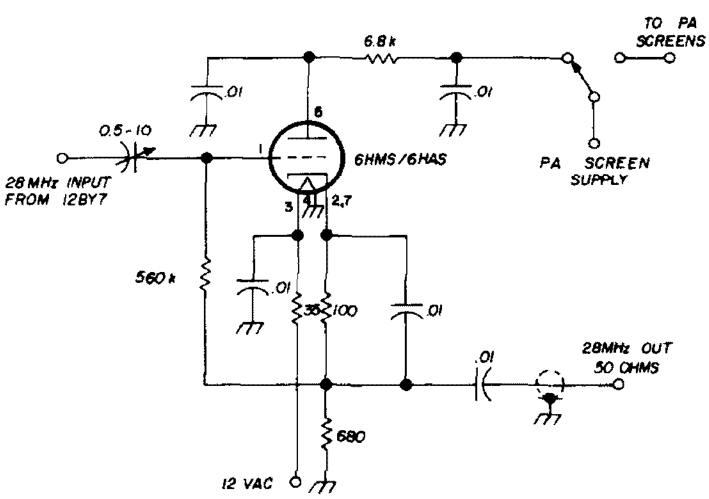



**fig. 4. Cathode follower for efficient power transfer from transceiver to transverter.**

It's possible to obtain the 28-MHz source form the transmitter or transceiver by (a) running the output into a dummy load and feeding a certain amount of energy through a small capacitor to the transverter, or (b) feeding the output to the transverter through an attenuator. These methods are inefficient and wasteful of power. Therefore, I've taken a new approach to the problem.

I use a type 6HM5/6HA5 tube in a cathode follower (**fig. 1**). This tube is very small, like the 6AK5. The circuit is installed on a little subchassis, which fits nicely in the 6JB6 input compartment of my TR4. The output bnc connector is at the rear, close to the 6JB6 output compartment, and is mounted with the switch on a small panel. When the switch is on, the 6JB6's are disabled, and no power is wasted.

The output impedance is given by $Z_{out} = 1/S = 1/0.02 = 50$ ohms, where S is the tube transconductance. With 135 V on the plate, this tube has a transconductance of 20k $\mu$mhos. This results in a normalized value of 50 ohms output impedance.

Because the input impedance of a cathode follower is high, only a very small amount of capacitance is needed to feed the 28-MHz signal from the driver stage, and no misalignment will occur.

Using this method I can obtain 1–10 volts of cw or ssb signal; it also works well on a-m. The stability of the vfo is the same as in the HF bands. The overall stability is determined by the crystal oscillator. I've been using this addition to my TR4 for two years on 144 and 432 MHz with very good reports. The idea can be used in other ssb equipment with equal results.

**Jacques Mainardi, F8MK**

## home-made heat sinks

Anyone who works with solid-state devices knows that when a large amount of power is applied to a transistor, a heat sink is required.

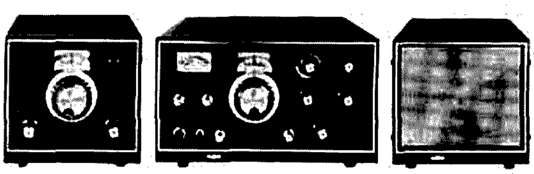
Many tv, fm, and other commercial broadcast stations use tubes such as the 4CX250 and similar types. In most cases these tubes aren't worth rebuilding when they go bad and can be had for the asking.

I found that by cutting the top section from the tube, plus a small amount of filing (and in some cases a little ingenuity), an excellent heat sink can be made for many power transistors. The convenience of having the heat sink around the transistor, rather than spread out along the chassis, can be realized.

**Greg Larsen, WA0WOZ**

# new products

## ssb,cw transceiver



Allied Radio has introduced a new five band ssb/cw transceiver that covers the range from 3.5 to 29.7 MHz in seven bandswitched ranges. The new transceiver features a solid-state vfo circuit with a linear tuning system that permits accurate readings to 1 kHz on all bands. The transceiver has built-in sidetone, vox,ptt, fast or slow agc, 25-kHz crystal calibrator, receiver incremental tuning and sharp-cutoff crystal filters, including a 500-Hz filter for cw work.

Power input is 160 watts from 3.5 to 21 MHz, and 120 watts on ten meters. Carrier and sideband suppression are rated at -40 dB. Receiver sensitivity is 0.5 $\mu$V for 10-dB signal-to-noise ratio (3.5 to 21 MHz, 1.5 $\mu$V on 28 MHz). Selectivity on ssb is 2.4 kHz at -6 dB, 4.8 kHz at -60 dB. On cw the selectivity is 500 Hz at -6 dB and 1.5 kHz at -60 dB.

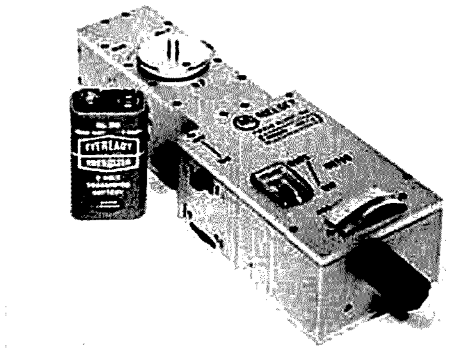
The Allied A-2517 transceiver is priced at $400. The A-2518 matching speaker/ac power supply is $99.95. A matching external solid-state vfo, the model A-2519, provides increased versatility by allowing transmit and receive operation on different frequencies; price is $89.95. Allied products are available exclusively through Allied Radio Shack stores, or by mail. For further information, write to Allied Radio Shack, 100 N. Western Avenue, Chicago, Illinois 60680.

## transistor manual

The new edition of the RCA "Transistor, Thyristor and Diode Manual" includes the latest available information on basic technology, operating principles, characteristics and ratings, applications and test of RCA semiconductors. This new manual is 20 percent larger than its predecessor and continues as an authoritative reference on bipolar transistors. In addition, it provides information on mos field-effect transistors, thyristors (scrs, triacs and diacs), silicon rectifiers and other types of solid-state devices. Definitive data are given for more than 900 different semiconductor devices; comprehensive data and design curves for transistors and thyristors are provided. In addition, tabular data are given for silicon rectifiers, other semiconductor diodes and discontinued transistor types.

In the circuits section of this manual schematic diagrams, detailed parts lists and descriptive writeups are provided for 38 practical circuits. Most interesting to the amateur radio experimenter are a mosfet preamplifier for 6, 10 and 15 meters, a two-meter converter, a stable vfo (3.5–4.0, 5.0–5.5 or 8.0–9.0 MHz output), microphone preamplifier, 40-watt 50-MHz transmitter with load mismatch protection, transistor dip meter and an electronic keyer. Other applications include fm tuners, an fm stereo multiplex demodulator, hi-fi amplifiers, power supplies and voltage regulators, battery chargers, an electronic heat control unit, light flashers and dimmers, and several digital circuits. 656 pages. $2.50 from your local RCA distributor; ask for Technical Series SC-14.

## uhf wave dip meter



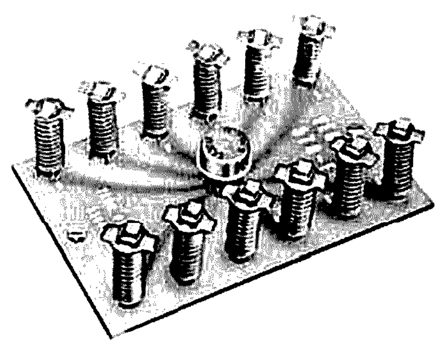
A new solid-state uhf wave/dip meter and marker oscillator has been introduced by the Melsey Corporation. This new instrument provides continuous tuning from 400 to 1150 MHz. Frequency readout—by the use of a 30-inch steel tape—is better than 1%. Design features include a battery-operated transistor oscillator in a cavity configuration that is tuned by a precision glass-invar capacitor. The instrument has an outside coupling loop for general use, and a miniature coaxial receptacle for direct connection with 50-ohm coax cable. The connector assembly can be rotated to obtain variable attenuation (30dB minimum)

In addition, special applications are possible with modifications developed by the manufacturer. With a minor adjustment the output of the unit can be increases to function as a local oscillator for uhf mixers. It can also be tracked to an rf circuit, or used as an fm, a-m or pulse modulated signal generator or target transmitter. The SN-2 wave/dip meter is $185 from Melsey Corporation, 202 Carle Road, Carle Place, L. I. New York 11514

## fet applications handbook

If you're looking for practical design data on field-effect transistors, this new expanded 2nd edition by Jerome Eimbinder, managing editor, *EEE Magazine,* contains nearly 25% more material than the previous volume. The in-depth information furnished by editor Eimbinder will be of immediate value to anyone looking for new ideas and unique fet circuit applications, including many basic fet circuit descriptions. Contents include introduction to the fet and basic fet characteristics, biasing fet stages, fets as oscillators, low-noise audio preamplifiers, source followers, phase splitters and switches. Also included are fet measurements, the photo fet, mosfet biasing techniques and fets as voltage-controlled resistors. The appendix includes often-needed data and charts arranged to serve as a convenient quick-reference source. 352 pages. $14.95 from Tab Books, Blue Ridge Summit, Pennsylvania 17214.

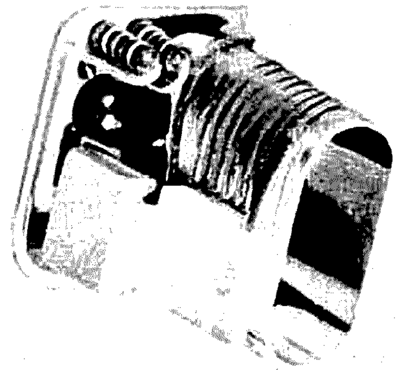
## ic breadoard socket



Vector Electronic Company has announced a new breadboarding socket that is designed for 12-lead TO-5 integrated circuits. The device consists of an epoxy-glass wafer with a 12-pin socket, the tabs of which have been soldered to two adjacent rows of solderless *Springclip* terminals. The board is furnished with two pins on the bottom that may be press fitted into pre-punched terminal board with 3/32 inch holes such as AA-pattern Vectorboard.

This new breadboard socket simplified integrated-circuit experiments because as many as four solderless connections can be made quickly to any terminal pin, and the user may use as many ICs as he needs by simply using additional breadboard sockets. The price of the 570F IC socket is $3.95 and may be ordered from the manufacturer, Vector Electronic Company, Inc., 12460 Gladstone Avenue, Sylmar, California 91342.

## ambient compensator



Isotemp Research, Inc. has recently announced an ambient temperature compensator for quartz crystals that uses a proportional solid-state control circuit to maintain crystal temperature within ±0.05° C at constant ambient temperature and constant voltage supply. The set temperature of the unit is 75 ±2.5° C. Warmup time is approximately 6 minutes from -30° C. Maximum power demand is 4 watts; approximately 1¾ watts are required to maintain crystal temperature at -30° C ambient. Required supply voltage is 12.0 Vdc (6 Vdc to 24 Vdc are standard, 28 Vdc to 48 Vdc are available).

The model 1CL6P-2 ambient compensator is designed for one HC-6/U crystal holder, the model 2CL6P-2 holds two HC-6/U crystals. Models are also available for HC-13/U holders. The small quantity price for the 1CL6P-2 ambient compensator is $10.00. Order from Isotemp Research, Inc., 1216 Harris Street, Charlottesville, Virginia 22901. Isotemp Research specializes in the design and manufacture of temperature-control sub-assemblies for electronic equipment and offers a variety of proportional-control crystal ovens in very small packages.

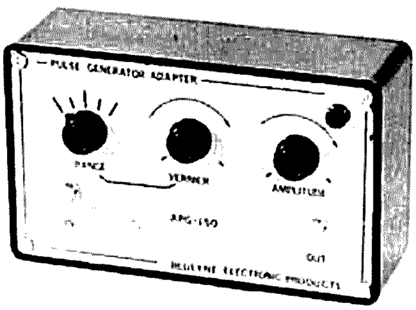
## two-meter amplifier

The new solid-state two meter power amplifier from Dynamic Communications puts out 10 watts with a maximum of 20 mW drive. The 101-500 power amplifier operates with a 12-volt power supply and is an ideal booster for 2-meter fm walkie-talkies or as a mobile final amplifier.

(With a 6-volt supply the amplifier puts out 4 to 5 watts.)

The DyComm power amplifier is completely broadbanded and will operate anywhere between 143 and 149 MHz with no re-tuning. The unit measures 2x2x6", including the heat-sink enclosure. $59.95 from Dynamic Communications, Inc., 301 Broadway, Riviera Beach, Florida 33404.

## pulse-generator adapter



The all-new pulse-generator adapter from Blulyne Electronics Corporation is the answer for the experimenter who has a sine/square-wave generator but needs a high-speed laboratory-quality pulse generator. The new pulse generator allows you to calibrate your scope at high frequencies (50 MHz rise time), to test check amplifiers for frequency response, to control chopper circuits, or to test any electronic circuits requiring fast rise times. The generator features variable pulse width from 100 nanoseconds to 500 milliseconds (50% duty cycle maximum); pulse amplitude is variable from 0.6 to 10.0 volts, and rise and fall times are each less than 20 nanoseconds.

The unit may be used with any sine- or square-wave generator with an input from 1 Hz to 10 MHz; input impedance is 5000 ohms. Two models are available: The APG-150 with 50 ohms output impedance, and the APG-100 with 100 ohms output impedance. Price of the APG-150 is $49.95; the APG-100 is $39.95. For more information write to, Blulyne Electronics Corporation, 3 Sand Springs Road, Williamstown, Massachusetts 02167.

## principles of electronic technology

Most textbooks written for electronics fall into one of four categories: (1) Texts and manuals for engineers, usually pertaining to theory with a mathematical approach, generally at the upper college level. (2) Texts and laboratory manuals for training electronic technicians for maintenance of consumer or industrial devices. (3) Texts, handbooks and Q&A books for preparing the reader for an FCC exam, or to build and maintain amateur or commercial radio stations. (4) Hobby books presented for the casual hobbyist and basement experimenter.

These four classes of books sadly neglect a very important category: basic electronic theory, presented at a high school level of math but with attention to detail that would do honor to an engineering text. Carl B. Weick's new book, "Principles of Electronic Technology," fills the gap.

Weick has done an admirable job with his book. It goes into great depth and detail in explaining how and why the basic circuit elements (R, C and L) function. A similar treatment is given to the behavior of fundamental ac and dc circuits. Electronic devices and wiring diagrams are left to other texts. What's unique about "Principles of Electronic Technology" is the thoroughness with which it prepares the student for progression to other levels of electronic technology, whether it be maintenance, engineering, or being a true *amateur* of radio. It is best suited for study in conjunction with an organized class. Like many other books, though, a home reader with real determination can master the text quite well. This, as always, requires reading, rereading, and thinking about each element presented, working out every problem and using every review question.

Don't be too ready to look down on basic electronic theory. There are few, other than practicing engineers and instructors who deal with such topics as daily routines, who truly have a compre-

hensive understanding of the basic subject. Why? Because engineers touch lightly on basics only as a hasty stepping-stone to the higher and more complex subjects. Once learned, the basics are quickly pushed to the back of their minds, to be recalled, if at all, only with studied effort.

Radio amateurs often scan only the surface of the basics, grasping only those facets we believe will be of use in passing the various grades of licenses examinations. We dig a bit deeper when we plan a construction project but then only in a narrow, specialized field of our immediate interest. If you feel your mastery of electronic theory is not as complete as you'd like, "Principles of Electronic Technology" may be just what you're looking for!

## power circuits manual

The latest edition of RCA's "Power Circuits Manual" has been updated and expanded to include the latest information on solid-state power devices. This new manual provides design information on a broad range of circuits that use power transistors, silicon rectifiers, and thyristors. In addition, it includes a brief introduction to semiconductor physics, as well as detailed descriptions of the construction, theory of operation, characteristics and circuit applications for each type of device.

The large comprehensive chapter on high-frequency rf power amplifiers covers the design of rf power amplifiers, matching networks, ssb transmitters, microwave amplifiers and oscillators, and frequency multipliers. Other chapters include rectifiers, power conversion, power regulation, thyristor ac-line voltage controls, and control and low-frequency power amplifiers. If your experimenting covers power-type semiconductors, you need this book on your workbench. Each topic is covered with the usual thoroughness that one expects from RCA. 448 pages. $2.00 from your local RCA distributor, as for Technical Series SP-51.

# short circuits

## bcd noise blanker

In the hot-carrier noise blanker circuit on page 17 of the October, 1969 issue of *ham radio* the 47k resistor connected to pin 4 of the second Nuvistor *should* be connected to pin 8. Pin 4 is connected directly to the .005 μF bypass capacitor.

## digital clock

Several parts values were omitted from the schematic diagrams of the digital clock on page 51 of the April, 1970 issue. R1 through R10 should be 20k for 170 Vdc operation of the Nixie tubes; R11 and R12 are each 270 ohms; C3 in the power supply should be 2000 μF. All gates are Fairchild μL914 or equivalent, and all flip-flops are μL923s.

Stafford Electronics advises the following price changes (first price is complete kit, price in parenthesis is for etched circuit board only): 12- or 24-hour clock, $145 ($13.50); 12- or 24-hour clock less seconds, $125 ($13.50); power supply/clock generator, $35 ($3.00); alarm circuit, $35 ($3.00); walnut cabinet, $30.

## solid-state power supplies

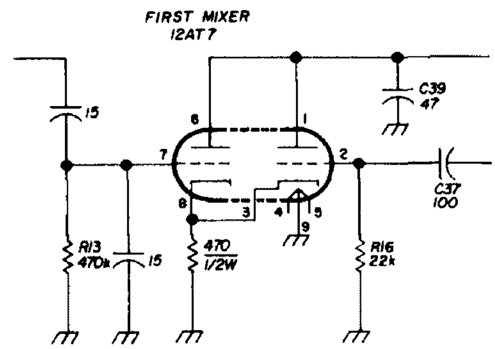
In fig. 21 of the "Survey of Solid-State Power Supplies" article, page 36, February, 1970, pin number 10 of the MC1460G voltage-regulator IC should be connected to the common bus.

## 32S1 modification

The wrong crystal was listed with the 32S1 modification that appeared in the *ham notebook* on page 82 of the December, 1969 issue. The correct Collins part number for the 457.550 kHz crystal is 290-8709-00. The parts department at Collins has been alerted to this error, so if you already ordered a crystal you should receive the correct frequency.

## 75A-4 modifications

There was an error in fig. **2B** of the 75A-4 modification article that appeared in the April, 1970 issue of *ham radio.* The The correct schematic is shown below.




## tilt-over mast

When calculating guy-wire tension (last item in fig. 7, page 48, February, 1970 *ham radio*) the wind loading at the top of the mast should be divided by the sin of angle B, *not* the tangent. The sine and tangent of 9.7° (example used in text) are numerically alike and no error of great magnitude exists, but for installations where angle B is much larger the error is large.

## 2-meter transmitting mixer

The E. F. Johnson Company has discontinued the type 189-253-5 PC butterfly capacitor specified for the 2-meter transmitting mixer featured in the April, 1969 issue of *ham radio* (fig. **5**, page 13). For C1 and C2, 8.5 butterflys, use Johnson 160-208. At C3, 10 pF butterfly, use Johnson 160-211; C4, 10 pF, use Johnson 160-104; C5, 20 pF, use Johnson 160-110. Capacitors C1 and C2 may be soldered upside down to the foil side of the printed-circuit board to the appropriate connection points.

75 cents

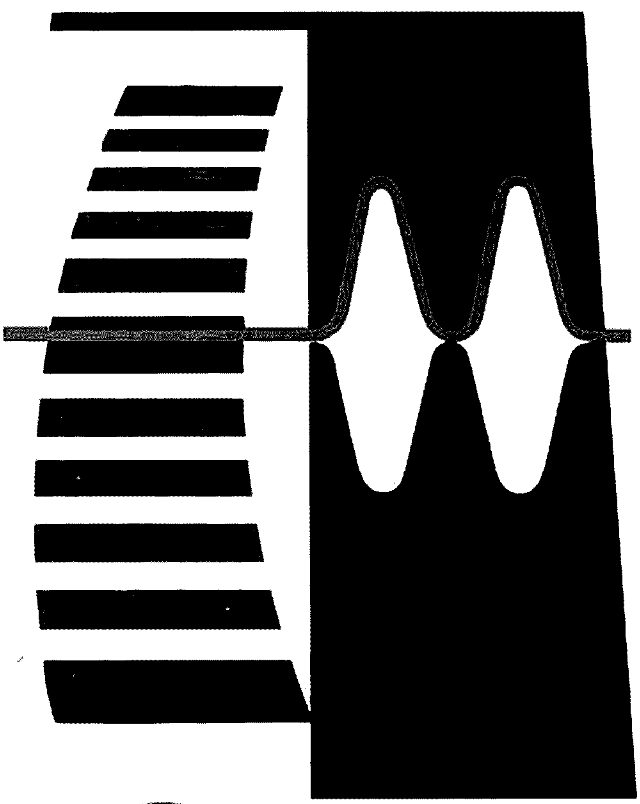
focus on communications technology . . .

# ham radio

magazine

OCTOBER 1970

## an swr meter



## for accurate rf power measurements

*this month*

October 1970
volume 3, number 10



**offices**

Greenville, New Hampshire 03048
Telephone 603-878-1441

**ham radio** magazine is
published monthly by
Communications Technology Inc.
Greenville, New Hampshire 03048

Subscription rates, world wide:
one year, $6.00, three years, $12.00.
Second class postage
paid at Greenville, N. H. 03048
and at additional mailing offices

Foreign subscription agents:
United Kingdom:
Radio Society of Great Britain,
35 Doughty Street, London WC1, England.

All European countries:
Eskil Persson, SM5CJP, Frotunagrand 1,
19400 Upplands Vasby, Sweden.

African continent:
Holland Radio, 143 Greenway,
Greenside, Johannesburg,
Republic of South Africa



Printed by Wellesley Press, Inc.
Wellesley, Massachusetts 02181, U.S.A.

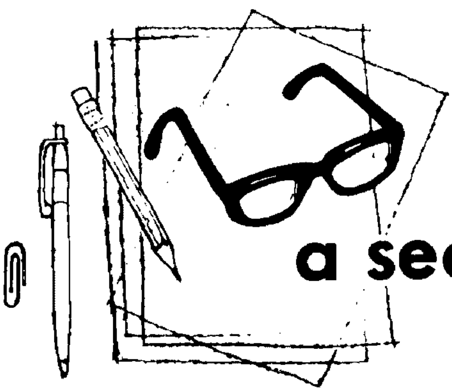
**ham radio** is available to the blind
and physically handicapped on magnetic tape
from Science for the Blind,
221 Rock Hill Road, Bala Cynwyd,
Pennsylvania 19440.
Microfilm copies of current
and back issues are available
from University Microfilms,
Ann Arbor, Michigan 48103.

Postmaster: Please send form 3579 to
**ham radio** magazine, Greenville,
New Hampshire
03048

hr

# ham radio magazine

## contents



# a second look

by jim fisk

**High-frequency radio communications** has always been at the mercy of the sun, with its sporadic cataclysmic bursts of electromagnetic radiation that convulse the ionosphere. The mighty sun will always dictate the condition of the ionosphere, but the unpredictability of solar disturbances is probably as disconcerting as the disturbances themselves. A new solar-activity monitoring system, installed this summer by Cambridge Laboratories, may end the unpredictability once and for all.

Cambridge Labs tried once before to develop a solar-activity prediction system, but it didn't work out; primarily because of the large number of errors, reporting inconsistencies and variations in instrument calibration.

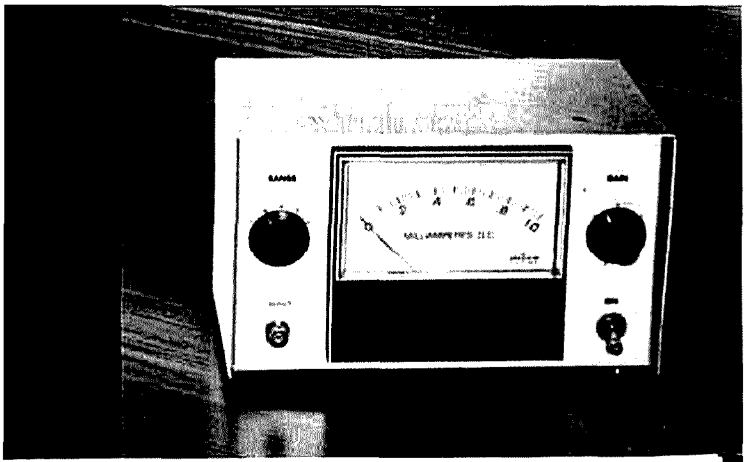
The new program, however, is different. While the observation stations were under development, a parallel program was initiated to establish uniform reporting and observing techniques, as well as carefully calibrated equipment that would provide standardized information.

The basis of the system is a huge network of sensors and data-processing equipment. The network consists of fifty ground-based optical and radio observatories all over the world. Each station is equipped with ionospheric sounders, radio receivers that measure galactic noise, neutron monitors and magnetometers. (Magnetometers are used to measure the intensity of the earth's magnetic field.) The ground stations transmit their data by teletype to a large computer complex at Offutt Air Force Base in Omaha. In addition, data from orbiting satellites is fed into the system, so the computer is up to date on the radiation level in space. The computer processes the data and forecasts changes that affect high-frequency communications.

Although the program is being conducted primarily for the benefit of the Air Force, amateurs can expect some fallout in the form of improved propagation notices from WWV.

In the future, when WWV broadcasts an N7 propagation forecast, you should be able to expect pretty good, undisturbed DX conditions with some degree of certainty. There are bound to be some anomalies until they shake all the bugs out, but hopefully some of the sorcery and ambiguity of propagation forecasting will soon be a thing of the past – a thing to reminisce about, like the time 9Z1AA dropped into the solar noise just as he answered your call.

**Jim Fisk, W1DTY**
**editor**



# the swr meter

Not to be confused with swr bridges, this laboratory-quality instrument is essential for serious uhf work

Bob Melvin, W6VSV, 113 El Camino Real, Millbrae, California 94030

As the experimenter progresses through vhf into uhf and the microwave regions, it becomes evident that instrumentation is required to assure optimum performance of equipment and antenna systems. For the most part, the amateur suffers under the handicap of being unable to adequately measure loss or gain at high frequencies. For this reason many can only guess the actual gain of their antenna or rf preamplifier, or the actual loss in their transmission line or rf filters.

Very little has appeared in amateur publications to assist the experimenter to assign meaningful loss or gain values to the various components of his systems. For example, crosstalk in antenna relays or loss in baluns are unmeasurable with most amateur equipment. Only by knowing the performance of the various components in the complete system can one obtain the overall performance desired.

The amateur often bases measurements of gain or loss on such unreliable devices as a receiver S-meter or a through-line watt-meter. These indicators along with pilot lamps, rf ammeters, and many "swr bridges" leave much to be desired at frequencies above 100 MHz.

This article describes the construction and operation of one of the most useful measuring tools for the vhf-uhf experimenter. No calibration instruments are required—all measurements are simply derived from the indication on a 0-1 dc milliammeter. With this instrument the amateur can determine meaningful decibel loss or gain figures previously unmeasurable, including gain in antennas or preamps, loss in attenuators or transmission lines, and crosstalk in coaxial relays.

## the swr meter

Drawing upon the experience of the microwave industry,[1] it is found that for many years measurements in the vhf through microwave regions have been

The swr meter has two major features:

1. It indicates power ratio directly.

2. It requires no calibration from an rf standard.

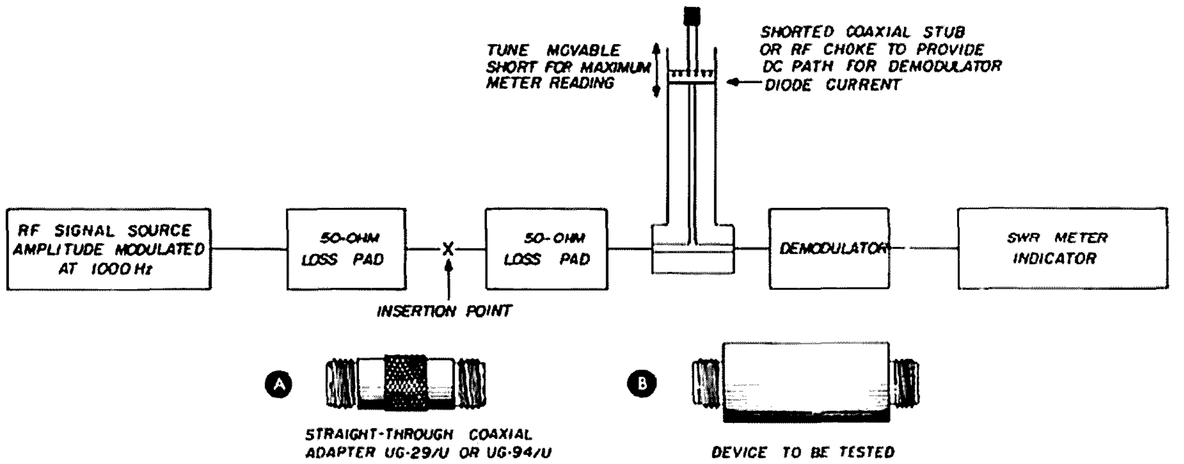



**fig. 1. Instrumentation for measuring power loss. Meter is first adjusted for full-scale deflection with straight-through adapter inserted at point X; then device to be checked is inserted.**

made with a laboratory instrument termed a vswr indicator or simply swr meter. This instrument has far greater utility than its name would imply, as will be shown later.

The swr meter should not be confused with simple devices such as swr bridges, antennascopes, or other gadgets that base their calibration accuracy on the characteristics of their semiconductor diodes. The accuracy of such devices is doubtful at vhf and above, and each instrument must be individually calibrated from a known standard.

With good circuit design and careful use, swr-meter accuracy can be essentially that of the readout instrument. Power ratio may be converted to dB by applying a simple mathematical relationship or it may be taken directly from a graph.

For all loss or gain measurements a tone-modulated signal source, a demodulator, and an indicating meter are necessary. With this setup, power can be read out for transmission lines, filters, directional couplers, and other devices. Antenna gain may be measured indirectly over a suitable test range by comparing the

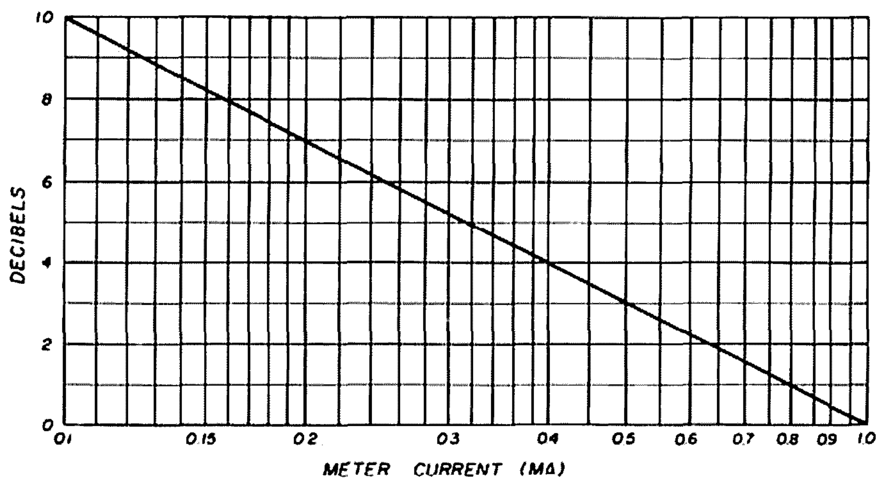



**fig. 2. Conversion chart for finding dB loss from meter indication.**

relative gain of the antenna under test with that of a carefully constructed reference antenna.

## theory of operation

The swr meter consists of a stable high-gain amplifier, tuned to an audio frequency, with a calibrated meter readout. The swr meter is designed to display power ratios when used with an rf demodulator (or detector) operating in the square-law region. Measurements are made on the demodulated audio-signal voltage output of the detector, as it is directly proportional to the incident rf signal power amplitude. By amplifying demodulated audio rather than rectified dc, stability and sensitivity are assured.

## using the swr meter

The test setup for rf loss measurements is shown in fig. 1. To make a reading, the instruments are connected as shown, and the swr meter gain is adjusted until the meter reads exactly full scale (1.0 mA or 0 dB). The circuit is opened at point X and the device to be tested is inserted. If the device presents a power loss at the signal frequency, the meter deflection will be less than full scale. The resultant fraction of full-scale reading represents the fraction of power flowing through the device under test. The loss can be converted to dB by

$$\text{loss in dB} = 10 \log_{10} \frac{1.0}{I} \qquad (1)$$

where

I = meter reading (mA)

This relationship has been plotted in fig. 2, which may be used with reasonable accuracy.

## slotted-line measurements

Although this article deals primarily with loss measurements, the swr meter can be used for its originally intended purpose—to measure vswr with a slotted line.[2,3] The slotted line is a mechanically precise section of transmission line, usually 50 ohms characteristic impedance, fitted with a movable rf detector probe (fig. 3). The probe samples the voltage field along the section of transmission line. The high gain of the swr meter is necessary to amplify the low-level of the sampled voltage in the line.

The combination of slotted line and swr meter not only provides greater accuracy than achievable with reflectometers or other instruments, but also indicates

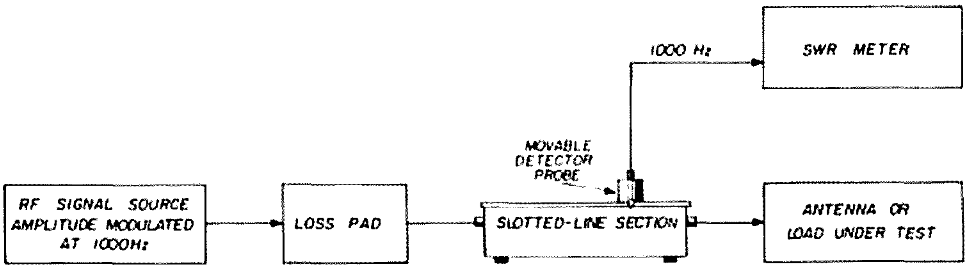
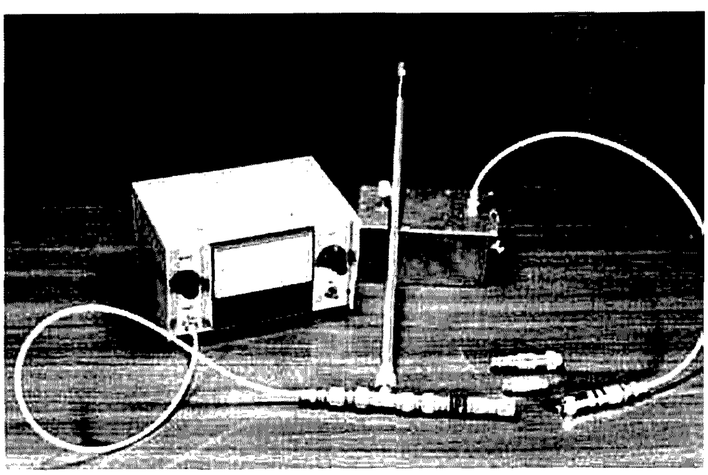



**fig. 3. Setup for measuring vswr with a slotted line.**



**Instrumentation for a complete microwave swr measurement setup. Shown are swr meter, signal source, and coaxial accessories.**

the position of the standing wave along the line. The location of the standing wave is necessary for solving transmission-line problems.

Measurements of vswr are made by first positioning the slotted-line probe carriage along the line until the point of maximum voltage is located. The swr meter gain is then adjusted to full scale (1 mA or 0 dB) at this point.

With the amplifier gain fixed, the probe carriage is then repositioned along the line to the point of minimum voltage or null. Vswr is the square root of the indicated power ratio:

$$\text{vswr}=\sqrt{1.0/1} \qquad (2)$$

where I is the meter reading in milliamperes.

This relationship is plotted in fig. 4. For vswr less than 3.16 curve A is used; for vswr greater than 3.16 swr-meter gain is increased by one 10-dB step and curve B is used.

## signal source

The rf-signal source may be a laboratory signal generator, although a readily constructed transistorized, crystal-controlled source will serve nicely. An output-power level of 1-10 mW is adequate for most measurements, including antenna range experiments. Special care should be used to assure low rf-harmonic content, rf-output amplitude stability, and constant modulation percentage. Minor variations of frequency do not normally cause loss of accuracy due to the relatively large bandwidth of the demodulator and device under test. The modulating tone, on the other hand, must be amplitude and frequency stable.

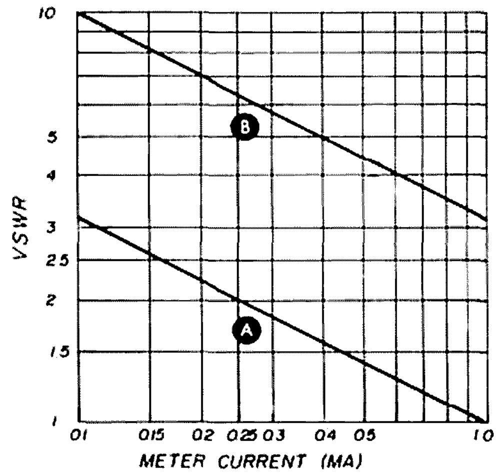
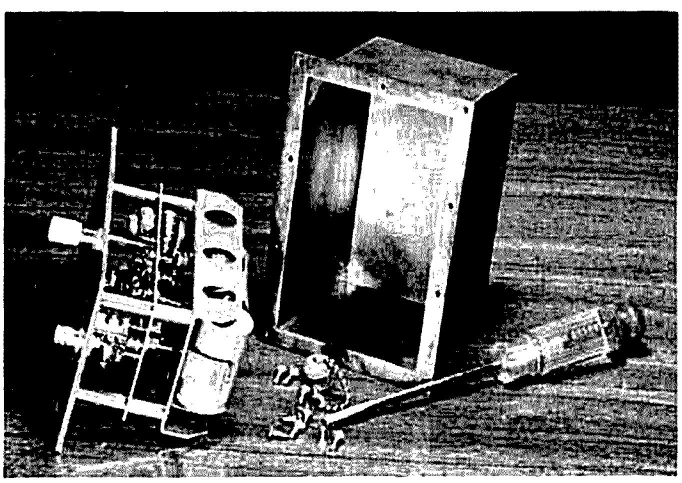



**fig. 4. Calibration curve for determining vswr as a function of meter readout.**

**Crystal-controlled 432-MHz signal source. Box seams were hard soldered to reduce rf leakage. Switch is actuated by a plastic rod through a brass tube soldered to panel.**



## demodulator

The demodulator, or detector, recovers the audio component from the amplitude-modulated signal. A coaxially mounted semiconductor diode, thermocouple, thermistor, or barretter are all capable of serving as a demodulator. For circuit simplicity and sensitivity, we have chosen to consider the semiconductor diode only.

Diodes display very accurate square-law characteristics at very low levels of incident rf energy. In the square-law region, a demodulator produces an audio output signal voltage directly proportional to its rf input power (or input voltage squared).

Point-contact diodes such as the 1N21 series in a suitable coaxial diode mount (see fig. 5) are useful for rf input levels producing less than 0.14 mV rms rectified audio output. At higher input levels, the diode characteristics deviate from the square law, thus reducing the accuracy of measurements.

A significant improvement in sensitivity can be realized by employing the

more advanced hot-carrier diode as a demodulator.[4] These diodes are now available for less than $1.00 each. The hot-carrier diode, provided with a small direct-current bias, will produce an 8- to 10-dB improvement in sensitivity over the 1N21. This additional sensitivity comes in handy for antenna-range work when using a low-powered signal source.

Fig. 6 illustrates a simple coaxial mount made of common coaxial connectors and a hot-carrier diode.

acceptable compromise between noise rejection and ease of operation.

The input circuit consists of an impedance-matching transformer and a network to provide dc bias for the hot-carrier diode demodulator. A 1:7 turns-ratio input transformer converts the low-impedance demodulator output to the optimum input impedance of the amplifier. A miniature output transformer, rated at 600:12 ohms, was selected. The turns ratio for this unit is 7.07:1. This transformer is a good choice, since it is much less susceptible to 60- or 120-Hz magnetic hum pickup than a higher-impedance transformer would be with the same turns ratio—example: 200:10k ohms.

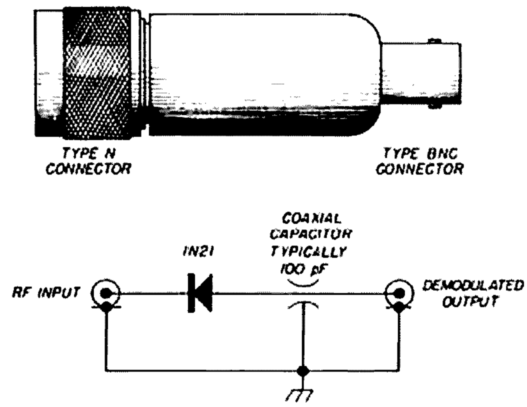



fig. 5. Microwave crystal demodulator using a 1N21 diode.

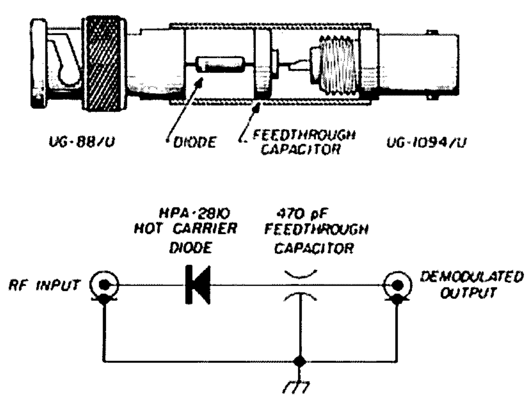



fig. 6. Hot carrier diode demodulator. Much higher sensitivity is possible using this device.

## amplifier-indicator

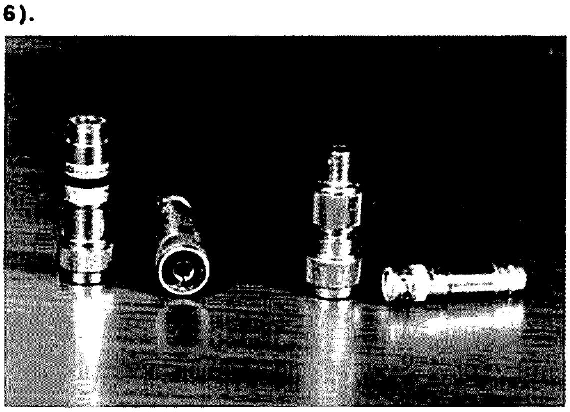
We have observed that semiconductor diodes behave in an accurately predictable manner at very low power levels. We can take advantage of these characteristics by using a high-gain amplifier and meter readout.

The amplifier must have sufficient voltage gain and power output to provide full-scale meter deflections for signals less than 1 $\mu$V. The circuit should be gain stable and linear for all signal levels. Thermal noise and hum effects should be suppressed.

The amplifier circuit of fig. 7 meets all conditions for gain, power output, stability and linearity. Amplifier linearity here implies that any increment of the 1000-Hz input signal will produce an exactly proportional increment in dc-meter current. Thermal noise and hum are suppressed to acceptable levels by limiting the amplifier 3-dB bandwidth to less than 50 Hz. This bandwidth is an

A forward bias of approximately 30 $\mu$A dc is necessary when using a hot-carrier diode demodulator. The bias may be disabled when using a 1N21 or 1N23,

50-ohm coaxial components. From left to right, two attenuator pads, commercial crystal holder (fig. 5), and homemade demodulator (see fig. 6).



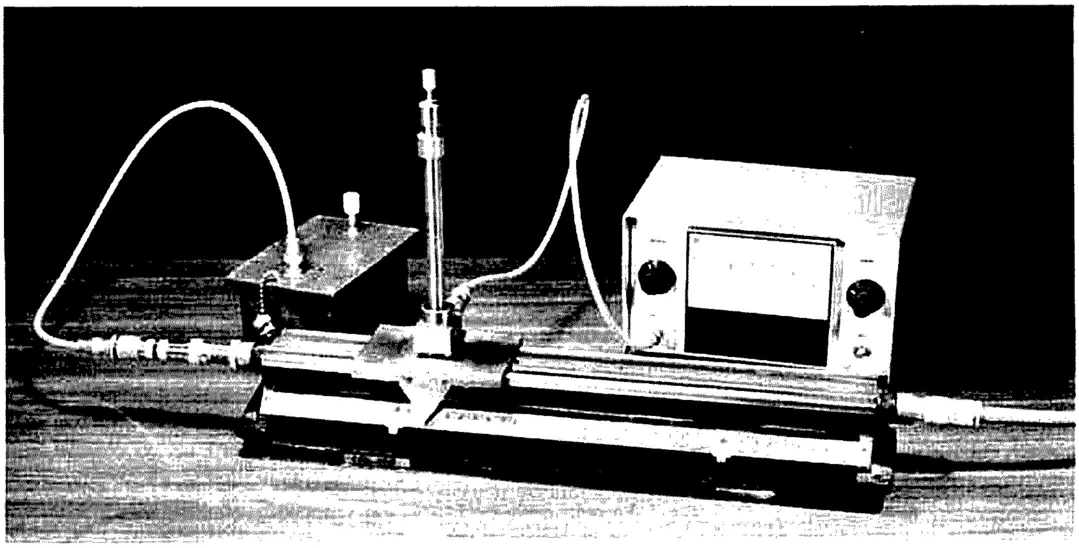
since point-contact diodes generate a great deal of noise with dc bias current.

Following the step-up transformer is the first section of the range switch. This switch attenuates the gain of the amplifier in exactly 20-dB steps. Because the instrument is to be used only with square-law detectors, each attenuation step is marked as 10 dB. Attenuator resistors are of the metal film type, 1% tolerance.

A four-stage, direct-coupled, high-gain transistor preamplifier contains coarse and fine gain controls in its negative feedback loop. The second section of the range switch at the output of the preamplifier prevents overloading of any stage over the entire range of inputs to the swr meter.

A two-transistor feedback pair constitutes the second amplifier stage and drives a 1000-Hz LC filter consisting of an 88-mH telephone-loading toroid coil and two Mylar capacitors, C1 and C2. Due to the very low output impedance of this amplifier stage, capacitors C1 and C2 are effectively in parallel and resonate the coil near 1000 Hz. The values of these capacitors are proportioned as shown to provide the required passband shape factor. If it is desired to peak the amplifier response at exactly 1000 Hz, a few turns will have to be removed from the toroid. Of greater importance is the matching or tuning of the modulating frequency of the signal source to the LC filter in the amplifier.

Four transistors in a direct-coupled configuration provide additional gain, phase-inversion, and complementary symmetry output. The 1000-Hz ac amplifier output is available for external use, if desired, but is not directly used to deflect the meter. It is the dc current to the class-B complementary symmetry stage that drives the meter, thus accomplishing the ac-to-dc conversion. The large amount of feedback in this stage reduces the closed-loop voltage gain to near unity. A 0-1 dc milliammeter is a good choice for indicator, as it is rugged, inexpensive, and readily available.

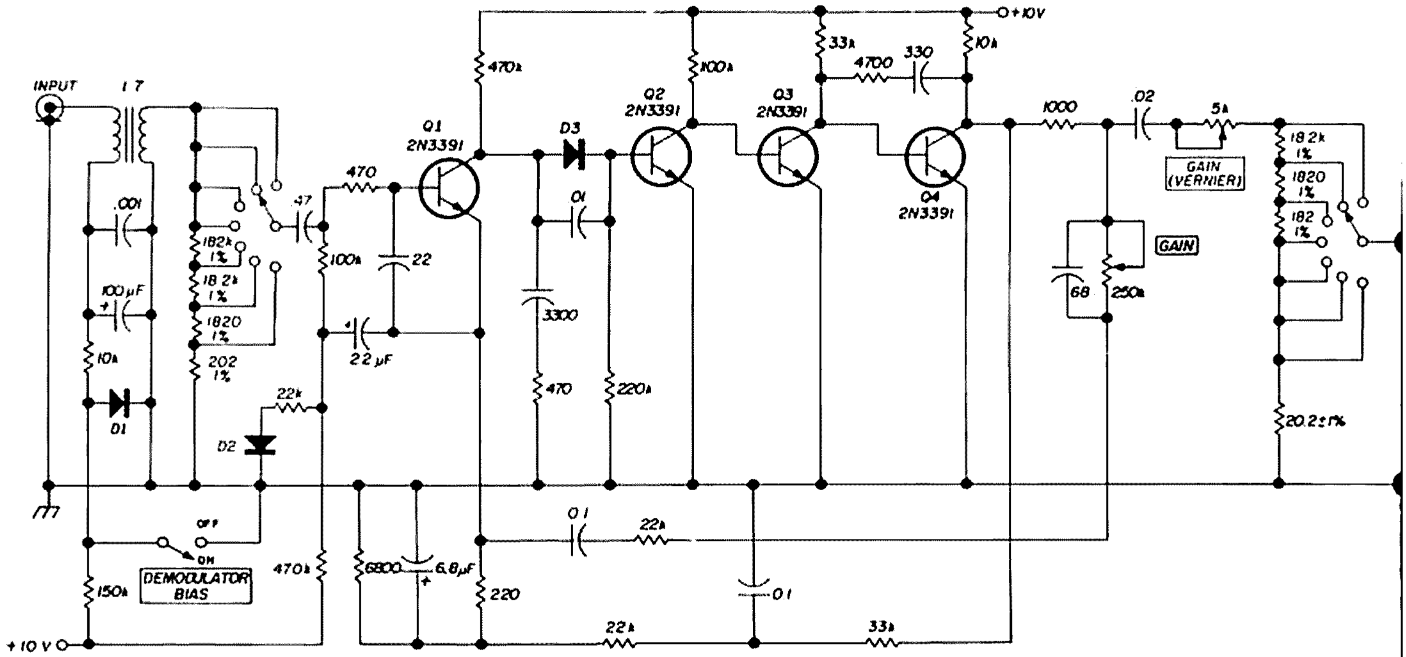


**Using the swr meter with a slotted line.**

Other circuitry includes a simple voltage-regulated power supply to maintain constant voltage to the amplifier as the battery voltage declines throughout its useful life.

## construction

Components are mounted on perforated board for simplest construction. Care was given to avoid ground loops by having only one ground connection to the metal cabinet at the BNC input jack. A bus bar ground wire, with each ground made in the same sequence as shown in




D1, D2, D3 • SILICON DIODE, 25V OR GREATER

**fig. 7. Schematic of complete swr meter for microwave measurements. Sufficient gain, output, stability, and linearity are provided for input signals less than 1 $\mu$V.**

the diagram, will eliminate ground loops. The precision range resistors are mounted directly on the range switch.

## instrument performance

The completed instrument meets the following specifications:

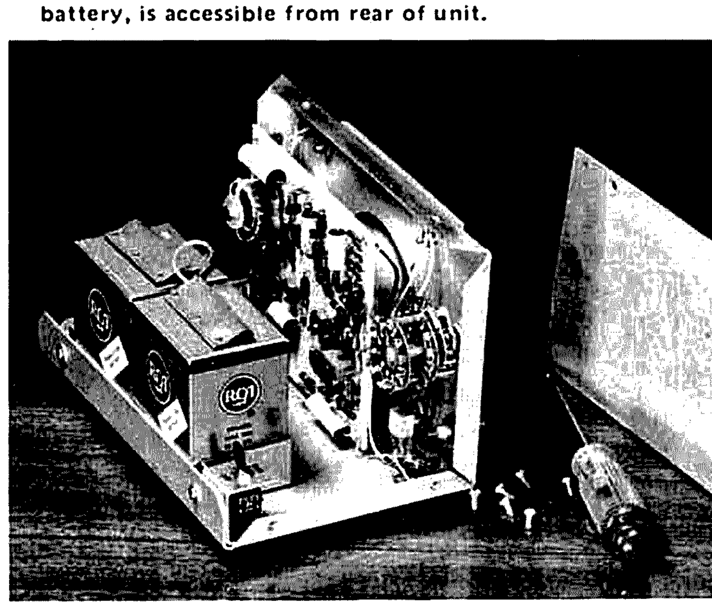
| | |
|---|---|
| sensitivity | less than 0.14 $\mu$V rms for full-scale meter deflection at maximum gain. |
| noise | greater than 10 dB below full scale at maximum gain. |
| gain | 70 dB in seven 10-dB steps using square-law demodulator. |
| bandwidth | 45 Hz at 3-dB down. |
| battery drain | 5 mA with no input signal, and 6 mA with input signal for full-scale meter deflection. |

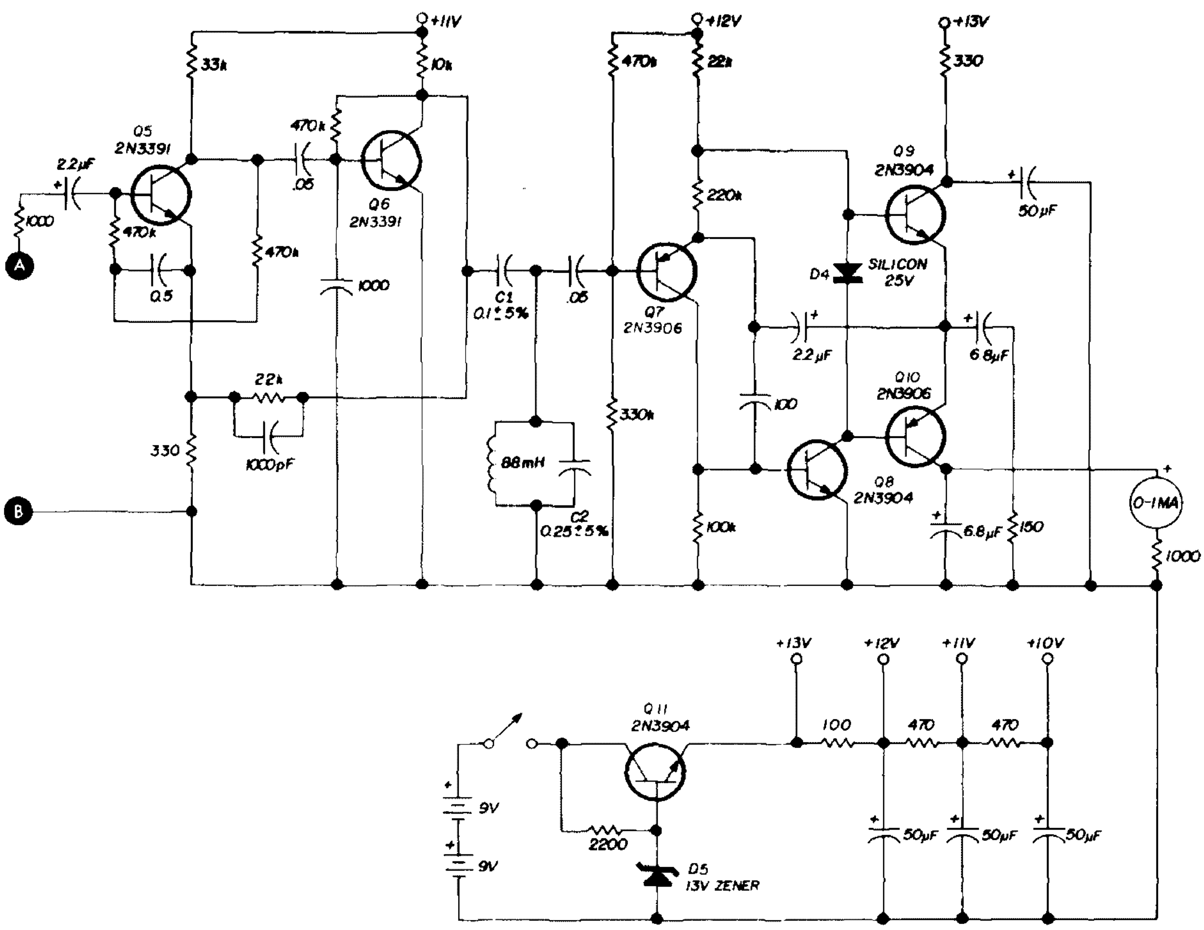
## error sources

A few precautions are necessary to realize full measurement accuracy. A significant source of error is caused by operating the demodulator at too great a level. Overload can be minimized by attenuating the signal source so that the swr meter gives on-scale deflection when operating on one of the three range-switch positions of greatest amplifier gain.

Measurement errors will be caused by any deviation from 50-ohm resistive

**The swr meter with cover removed. Circuit board is mounted on meter terminals. Demodulator bias on-off switch, mounted next to battery, is accessible from rear of unit.**





impedance of the signal source or detector loading.* A convenient solution is to insert a loss pad rated at 10 dB or greater after the signal source and another loss pad ahead of the detector mount. (See **fig. 1**.)

Additional error can result from any change of the dc resistance in the demodulator circuit caused by connecting or disconnecting various attenuator pads during tests. A solution to this problem is to connect a shorted quarter-wave stub, rf choke, or parallel-resonant circuit from demodulator rf input to ground. Such circuits display high shunt impedance for rf across the demodulator, while providing a low shunt impedance path for the 1000-Hz audio signal.

Good practice dictates special precautions when measuring attenuation greater than 20 dB or slotted line vswr greater than 10. For these measurements, a substitution technique using a previously calibrated loss pad can be of assistance, as will be shown in Example 3.

*An impedance of 50 ohms has been accepted as standard for coax lines, matched connectors, and test equipment. If it's necessary to use another impedance, all pads, connectors, and transmission lines must conform to the selected impedance.

## examples of operation

**Example 1.** Let us measure the loss of a length of 50-ohm coaxial cable at 432 MHz. The equipment setup of **fig. 1** is used with a 432-MHz signal source. The swr meter gain controls are carefully adjusted for full-scale meter deflection (1.00 mA or 0 dB). The junction (X) between the two loss pads is then opened, and the cable under test is inserted at this point. With the coaxial cable in the rf circuit, the meter indicates only 0.40 mA. This indication represents a transmitted power of 40 percent through the cable, or 60% loss in the cable at 432 MHz. Referring to eq. (1) we convert the meter reading to power loss in dB:

$$\text{Loss in dB} = 10 \log_{10} \frac{1.0}{0.40} = 3.98 \text{ dB}$$

**Fig. 2** may be used to simplify conver-

sion from meter reading (power loss) to dB.

**Example 2.** For our second example, suppose the unknown is a loss pad. When the pad is inserted in the line at (X), the meter reading is less than 0.10 mA. Without touching the coarse or fine gain controls, the range switch is switched in 10-dB steps until the meter reading falls between 0.10 and 1.00 mA. The number of 10-dB steps of gain change must be added to the reading.

In this example the meter reads 0.50 mA after the gain has been increased by one 10-dB step. The loss of our pad under test is:

$$\text{Loss in dB} = 10 + 10 \log_{10} \frac{1.0}{0.50} = 13.0 \text{ dB}$$

**Example 3.** Antenna-relay crosstalk is measured by first connecting a pad of known loss at (X) and adjusting the signal-source output and amplifier gain for exactly full scale. For this example we use a pad with 20-dB loss. The pad is now removed and the relay is connected at (X). (See **fig. 8.**)

With the relay actuated in the transmit mode, the meter reads less than 0.10 mA. With the coarse and fine gain controls fixed, the gain is increased by one 10-dB step of the range switch. The meter is then found to indicate 0.25 mA. The relay crosstalk loss is:

$$\text{Loss in dB} = 20 + 10 \log_{10} \frac{1.00}{0.25} = 36.0 \text{ dB}$$

**Example 4.** The gain of a receiver preamplifier is measured by first connecting it into the line at (X). The swr meter is then adjusted to exactly full scale. When the premaplifier is removed and a straight adapter substituted, the meter then reads 0.125 mA. From formula or chart we calculate the preamp gain to be 9.03 dB. By this procedure, we measure the loss when the preamplifier is removed from the circuit.

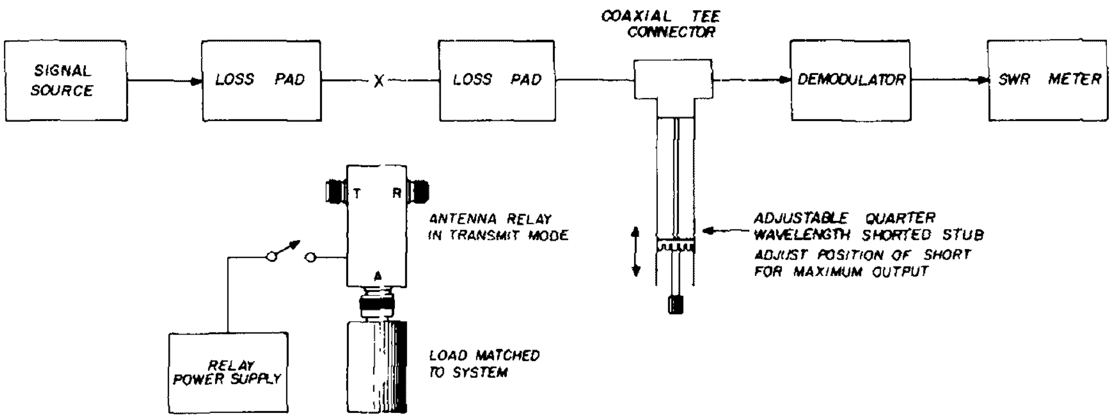



**fig. 8. Measurement arrangement for determining crosstalk in an antenna change-over relay. Meter is adjusted to full scale with loss-pad inserted at point X. Relay is then substituted and crosstalk is determined by adding pad loss to meter reading in dB.**

## conclusion

This article covers the construction and operation of an instrument of great utility for the serious experimenter in the uhf region. With careful construction the swr meter will perform as claimed. When used according to instructions given, the instrument will take the guesswork out of rf measurements at vhf and uhf.

### references

1. MIT Radiation Laboratory, "Microwave Techniques," U. S. Government Printing Office, NavShips 900, 028.
2. Hewlett-Packard Co., "Operating Manual, 415B or 415E."
3. Edward L. Ginzton, "Microwave Measurements," McGraw-Hill Book Co., New York.
4. Hewlett-Packard Co., "Hot Carrier Diode Video Detectors," Application Note 923.

**ham radio**


# the sideband minituner

Here's a pocket-sized direct-conversion receiver for 80 and 40 meters all for less than $25

Rick Littlefield, K1BQT/6, 209 Main Street, San Rafael, California 94901

This kitchen-tabletop project has appeal for everyone, from the beginner to the seasoned experimenter. It's perfect for cw practice, motel monitoring on trips, or for keeping an ear on your favorite net while the big rig is tied up on the DX bands. It can monitor your transmitter on ssb and cw, provide a signal source for tuning other receivers, and it can keep you in touch when power lines fail. If you are a QRP fan, it will provide you with a vfo and companion receiver for your little transmitter. Not to mention, of course, its value as a conversation piece when one of the locals drops over.

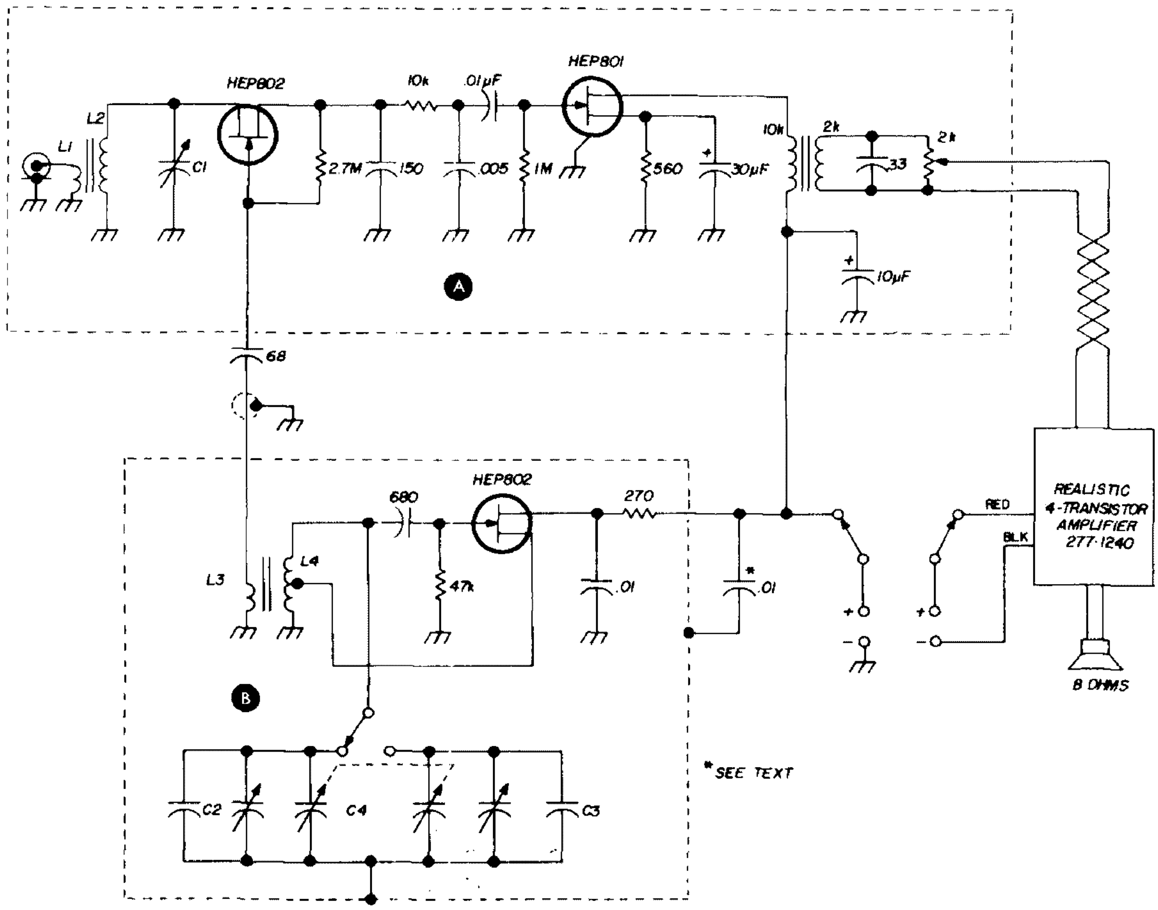
The inspiration for this project came from an article entitled "Direct Conversion—A Neglected Technique" by Wes Hayward and Dick Bingham. After working with this principle, I'm convinced that it should *not* be neglected, for there is no simpler or less-expensive approach to respectable ssb and cw reception.

## design

The heart of this receiver is an extremely simple product detector circuit (fig. 1). The preselected incoming signal is chopped at a rate determined by the vfo output frequency, with a resultant audio product appearing at the drain. Since the

vfo signal level at the gate is relatively high compared with the incoming signal level at the source, conduction time is short, and linearity is excellent. The output of the detector, after being fed through a low-pass filter, is amplified by a low-noise fet audio preamplifier. This single stage provides enough gain to drive one of the inexpensive imported audio modules. In this case, a Realistic 277-1240 100-mW unit was used. This unit requires an input of only 1 mW for full output, unmodified. With any reasonable antenna, the sensitivity of this circuit is far more than adequate.

Selectivity as well as sensitivity is determined within the audio range. The RC values shown provide reasonable selectivity for casual listening without sacrificing system gain. The Realistic 277-1240 audio board contains an 82k ohm feedback resistor between the output transformer secondary and the base of the transistor in the second stage. A 0.001-μF capacitor connected in series with the resistor does wonders in improv-




| | | | |
|---|---|---|---|
| **C1** | **365 pF miniature mica BC tuning capacitor** | **L1** | **3- or 4-turn link on L2** |
| **C2** | **470 pF small mica** | **L2** | **36 turns no. 30 enameled wire on .68 diameter toroid form** |
| **C3** | **140 pF small mica** | **L3** | **5-turn link on L4** |
| **C4** | **15/16 x 1-3/8 x 1-15/16 inches. 6.3—123.1 pF and 5.7—78.2 pF (Lafayette 32E11067 or equivalent)** | **L4** | **22 turns no. 22 enameled wire on .68 diameter toroid form, tapped 5 turns from ground end** |

**fig. 1. Schematic of the direct-conversion minituner cw/ssb receiver. The only circuits to build are the product detector/preamplifier, A, and the vfo, B.**

ing overall gain at voice frequencies. This combination causes the amplifier to be very frequency selective and decreases the interference problem between adjacent stations.

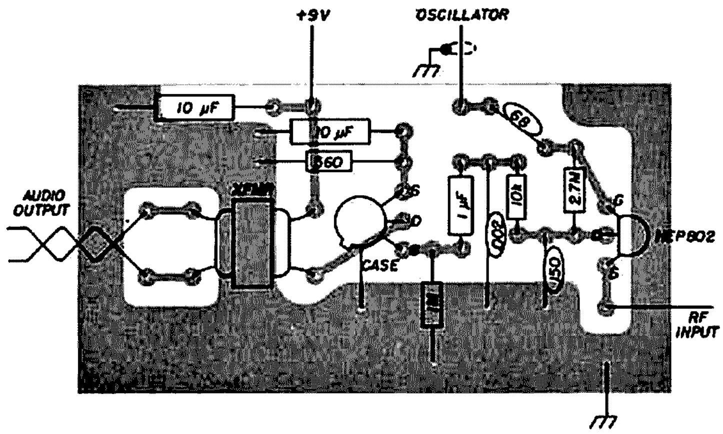



**fig. 2. Parts layout for the product-detector board. Components are shown as seen from the bottom of the board. Etched board is shown on the next page.**

## construction

The receiver consists of three small printed-circuit modules mounted in a 2 x 3 x 5-inch minibox. Since the audio amplifier is preconstructed, a good portion of the work is already done. The other two modules are extremely easy to lay out and construct. Fig. 2 and 3 show the board templates and component mounting.

Since the oscillator must maintain sideband isolation and stability, a few simple considerations are in order. If high-level vfo radiation reaches the rf input port of the detector via the antenna line and its tuned circuit, the linearity of the detector will be destroyed. Thus, the oscillator module must be well-shielded from the rest of the package. I accomplished this by building a separate box for the vfo from two-sided pc board. A less-elaborate shield could consist of a small minibox or an aluminum bracket. The power lead is bypassed at the vfo shield, and the vfo output lead is a length of small-diameter coaxial cable.

For the sake of stability, all frequency-critical capacitors are of the mica variety. The tuning capacitor, in this case a miniature air variable, has a ball-bearing rotor mount for smooth operation. All components on the module must be securely mounted, and all frequency-critical wire leads must be rigid. By following these few suggestions, you can expect quality reception from the finished project.

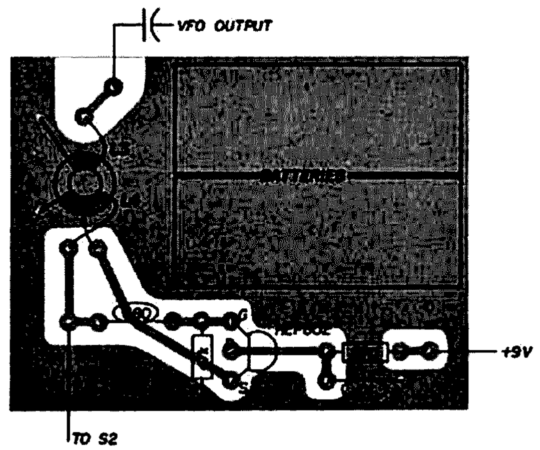
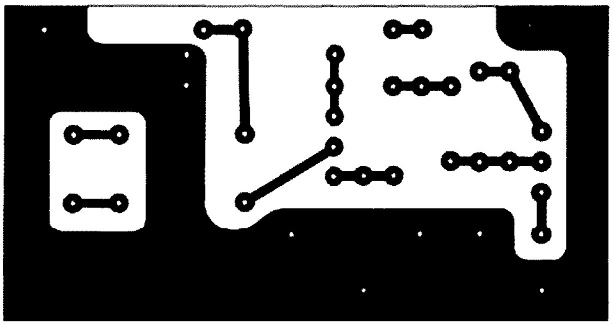



**fig. 3. Parts layout for the vfo circuit board. Components are viewed as seen from the bottom of the board. Etched board is shown on the next page.**

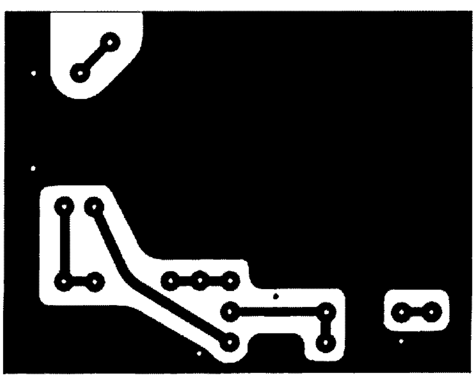
## adjustment

Tuneup is a relatively simple matter because the oscillator is the only section requiring initial adjustment. Since the same value of inductance is used for both the 80- and 40-meter bands, capacitance



is the only variable determining range and band placement. The variable- and fixed-capacitor values are thus considerably greater for 80 meters than for 40 meters. For the 2-section variable capacitor specified in the parts list, I left six rotor plates on the 80-meter section and only one on the 40-meter section. This provided about 300 kHz of coverage on each band. If you use one of the inexpensive imported 6:1 vernier drives, 300 kHz is the most you will be able to cover and still maintain smooth tuning. The fixed mica capacitor values should place the vfo output somewhere within the amateur bands, where you will be able to pick up the oscillator on your station receiver.

Minor fixed-capacitance substitutions



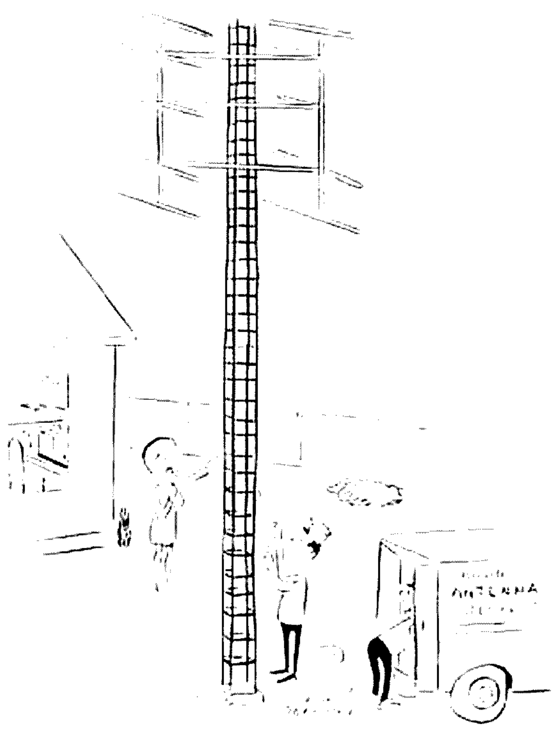
may be necessary to achieve the precise coverage you desire. Once you have the desired placement worked out on both bands, the station receiver can be used for an accurate final calibration. The tuned-rf input circuit covers 80 and 40 meters with enough overlap so that no inductance adjustment is necessary. Variable capacitor C1 peaks the incoming signal just as the antenna trimmer or preselector control on a commercial receiver. What could be simpler?

## conclusion

Despite its surprisingly good performance, this unit is not the ultimate in direct-conversion design. The possible avenues of exploration are many. For example, the Realistic 277-1577 audio board provides 350-mW output for 300-$\mu$V input at 500 ohms, and is well worth trying in place of the less-sensitive 100-mW module used here. Furthermore, a dual-gate mosfet product detector deserves consideration. And, with a stable vfo, this circuit should perform as well at 144 MHz as it does at 3.5 MHz. The possible arrangements are almost endless.

Nonetheless, and regardless of its potential for further development, the sideband minituner works very well as it is. In fact, you'll find it one of the most useful minibox projects you can build.

ham radio



"I think I'd like it better over by the fence after all."

# voltage-probe receiving antenna

A new two-inch-high electronic device that out-performs a 3-foot whip below 40 MHz

James Fisk, W1DTY

The voltage-probe antenna, or vpa, is a new approach to miniature high-performance broadband receiving antennas that exhibits low-noise characteristics and does not require tuning over the frequency range from 30 kHz to 50 MHz.

The vpa consists of an antenna probe and solid-state amplifying and matching circuitry designed to feed into 50-ohm coaxial line. The antenna probe itself consists of a rod two inches long, top loaded with a small disc as shown in the photo. The electronic circuitry is contained in the small cylinder below the antenna probe. Dc power, isolated from the incoming rf, is fed to the circuitry through the coaxial feedline.

## how it works

The vpa senses voltage from an incoming radio wave with its antenna probe, and the electronic circuitry acts as an impedance converter by transferring the input signal voltage from the highly reactive antenna probe to a 50-ohm resistive impedance level. The vpa performs much like a cathode follower, but with very high gain. The circuitry consists of an fet low-noise input stage, a buffer-driver stage and an output stage to match the probe to a 50-ohm transmission line.

The theory behind the vpa is based on the fact that when a quarter-wave monopole antenna above a ground plane is made infinitely short, its power gain

decreases only slightly from a theoretical maximum of 2.14 dB to 1.76 dB (over an isotropic). However, as the antenna gets shorter, antenna resistance decreases while the reactance increases. This is a familiar problem to the amateur who has worked with small 75-meter mobile antennas.

As an example, at 20 kHz, a 1-meter stub over a ground plane exhibits an effective resistance on one-millionth ohm, and capacitive reactance of 10 million ohms. Since these two components are effectively in series as shown in **fig. 1**, the voltage available to the amplifier is infinitesimally small.

The electronic circuitry used with the vpa does not match impedances in the usual sense. Instead, it makes the reactive component small in comparison to the very high input impedance of the fet amplifier, swamping out the reactive component. The result is that most of the input signal appears across the amplifier input terminals.

## performance

As pointed out by K6MIO,[1] the ability of an antenna to extract a radio signal from space is dependent upon the antenna's effective aperture or capture area. This capture area may be used as a basis to compute the antenna's gain as compared to an isotropic. The gain of the vpa, when placed 42 inches over a ground plane, is −10 dB, and is flat from 30 kHz to over 50 MHz. As a point of comparison, a quarter-wave whip (matched to the receiver) exhibits 2.14 dB gain. However, the whip is a high-Q device, and performance deteriorates rapidly as you deviate from its resonant frequency.

The results of a comparison test between the miniature voltage-probe antenna and a 13-foot, 4-inch whip are plotted in **fig. 2**. For these tests, conducted by the Kollmorgen Corporation, both antennas were located side-by-side on a 50 x 80-foot metallic ground plane. The whip sat on a 9-inch pedestal *in contact* with the ground plane; the vpa was on a 42-inch post, also in contact with the ground plane. The outputs, brought out through 50-ohm coax, were compared on the same receiver; a precision variable attenuator was used to measure differences in signal level.

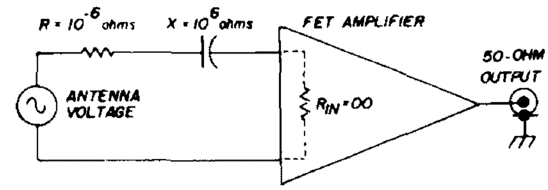
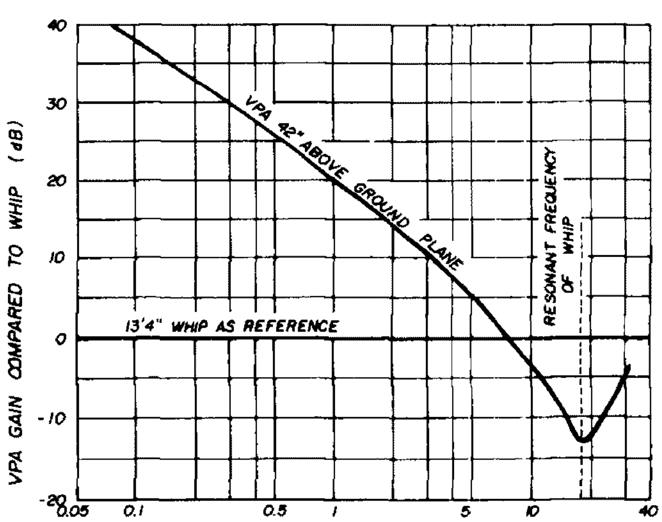



**fig. 1. Equivalent circuit for a miniature 1-meter-long stub antenna at 20 kHz.**

**fig. 2. Performance of the voltage-probe antenna as compared to a 13' 4" whip.**




Performance of the vpa can be improved by elevating it further over the ground plane. This is because of signal pickup by the output coaxial cable. Although vpa gain is flat to 50 MHz, the effective gain over a whip antenna increases substantially with decreasing frequencies. The vpa shown in the photo, for example, will out-perform a 3-foot whip antenna at all frequencies below 40 MHz.

### references


1. James Kennedy, K6MIO, "Antennas and Capture Area," *ham radio,* November, 1969, page 42.
2. Robert Fischer, "Voltage Probe Antenna," Kollmorgen Corporation report ER848.1, 29 January 1968, Kollmorgen Corporation, Northampton, Massachusetts 01060.


**ham radio**

# designing with ic voltage regulators

Inherent device constraints are analyzed, and design examples are given to obtain optimum regulator performance

Don Steinbach, W7FLC, 1531 McKinnon Court, San Jose, California 95130

**Some of the load-handling capability** of an IC voltage regulator may be sacrificed if a less-than-optimum system design approach is used. This article describes a design procedure that assures full realization of the regulator's load-handling capability and compliance with all its ratings. Design examples, using a typical IC voltage regulator, are included.

The cost of a monolithic IC voltage regulator is usually less than that of a voltage regulator of comparable performance constructed from discrete components. In return for this reduced cost (and increased convenience) the designer is confronted with a set of regulator operating constraints, imposed by the IC manufacturer, over which the designer has little control.

The voltage regulator IC is a complete electronic circuit block, and certain package terminal conditions must be met regardless of the regulator's relationship to the rest of the system. These are:

1. The minimum regulator input voltage

2. The maximum regulator input voltage

3. The maximum regulator output current.

Two more constraints are added when the voltage regulator becomes part of a voltage-regulated power supply system:

4. The minimum voltage drop across the regulator

5. The maximum allowable regulator power dissipation

Since the IC operating parameters and the power requirements of the load are predefined, the volt-ampere characteristic of the unregulated voltage source is the *only manipulatable system variable.* The boundaries represented by the preceding constraints can be plotted on a graph of regulator input voltage vs regulator output current, and will enclose a region that contains all the allowable combinations of unregulated input voltage and regulator output current. Obviously, the volt-ampere characteristic of the unregulated

power supply must lie within the boundaries of this permissible operating region. The mechanics of this technique are demonstrated later in the article.

## device limitations

The *maximum input voltage* that may be applied to an IC voltage regulator is restricted by the IC secondary breakdown limitation and, when current flows, the maximum allowable junction temperature. The *maximum current* that may pass through a voltage regulator IC is restricted by the current-carrying capacity of the interconnections within the IC, the maximum allowable junction temperature, and the voltage applied to the regulator when the second-breakdown region is approached. Power dissipation limits will normally be exceeded before second breakdown is reached with increasing current—unless, of course, the case temperature is prevented from rising.

The *minimum input voltage* applied to an IC regulator must be adequate to maintain appropriate bias currents through the internal zener diodes. Furthermore, a minimum input-output voltage differential must be maintained to assure that the series pass regulating elements retain control over the output voltage. The greater of these two minimum input voltage restrictions must always be satisfied. The *maximum allowable power dissipation* is limited by the maximum allowable junction temperature and the thermal resistances from junction to ambient.

The numerical values and package pin connections cited in the remainder of this article apply specifically to the Motorola MC1469R connected as shown in **fig. 1**. The comments and procedures are, however, equally applicable to all other voltage regulator ICs.

## typical regulator circuit

A complete voltage regulator circuit using a typical IC voltage regulator (the Motorola MC1469R) appears in **fig. 1**. The unregulated input voltage is applied to pin 3 of the MC1469R, and the IC output voltage appears at pin 1. Capacitors C1, C2, C3 provide circuit compensation, noise filtering, and assure output-voltage stability.

The 2N706 transistor, Q1, and current-sensing resistor, $R_{sc}$, limit (via pin 4) the maximum output current of the regulator to that which causes a voltage drop of 0.6 volt across $R_{sc}$. Thus, the short-circuit current, $I_{sc}$, is defined by $R_{sc}$ and is determined from the expression

$$I_{sc} = \frac{0.6}{R_{sc}}$$

where $I_{sc}$ = the approximate short-circuit current limit in amperes, and $R_{sc}$ = the resistance in ohms of the current-sensing resistor.

The regulated and current-limited output voltage, $V_o$, appears at the emitter of Q1. This voltage is sensed by pin 5, and the voltage at pin 1 varies, as required, to keep the voltage at pin 5 constant. The constant voltage required at pin 5 is a function of R1 and R2, and R1 is determined from

$$R1 = 2V_o - 7$$

where $V_o$ = the desired regulated output voltage in volts, and R1 = the resistance (in k ohms) of the voltage-setting resistor.

Pin 2 of the MC1469R is the shutdown control. A positive voltage applied to pin 2 causes the regulator output voltage to drop to zero and reduces the regulator input current to a few-hundred microamperes. Aside from its obvious use as a regulator on/off control, the shutdown control automatically limits the maximum IC chip temperature if a constant positive reference voltage is applied to pin 2. The voltage required to shut down the regulator is a function of the chip temperature:

$$V_{pin\ 2} = 1.38 - 3.4 \times 10^{-3}\ (T_j - 25^{\circ}C)$$

where $V_{pin\ 2}$ = the approximate voltage in volts that must be applied to pin 2 to shut down the regulator at a given junction temperature, $T_j$, and $T_j$ = the junction temperature in °C at which shutdown will occur. Since the shutdown voltage decreases approximately linearly

at the rate of 3.4 millivolts/°C, the shutdown voltage under a particular set of operating conditions can be determined experimentally and its value used to calculate the chip temperature rise. This technique has the advantage of not disturbing the thermal characteristics of the IC through the attachment of an external temperature-measuring device.

The continuous load current from the MC1469R must not exceed 500 mA. It will withstand a maximum input voltage from pin 3 to ground of 35 Vdc; the minimum input voltage must be 9.0 Vdc or 3.0 Vdc higher than the voltage at pin 1, whichever is greater.

The voltage at pin 1, $V_{pin\ 1}$, can be determined only after the regulated output voltage, $V_o$, and $R_{sc}$ have been specified. The equation:

$$V_{pin\ 1} = V_o + I_L\ (R_{sc})$$

where $I_L$ = the load current in amperes, is valid for all $I_L$ from 0.0 to $I_{sc}$. Note that $V_{pin\ 1}$ increases linearly from $V_o$ at $I_L$ = 0 to $(V_o + 0.6)$ at $I_L = I_{sc}$.

The MC1469R can dissipate 3 watts at an ambient temperature of 25°C without a separate heat sink. The actual power dissipation is calculated from the voltage drop across the IC and the current into the IC, and is closely approximated by

$$P_D \text{ actual} = (V_{pin\ 3} - V_{pin\ 1})\ I_{pin\ 3}$$

where P is in watts, V is in volts, and I is in amperes.

The voltages and currents are defined in **fig. 1**. If the MC1469R is operated with no heat sink at ambient temperatures other than 25°C, the maximum allowable power dissipation can be computed by

$$P_{Dmax} = \frac{150 - Ta}{41.6}$$

where $P_{Dmax}$ = the maximum allowable device power dissipation in watts at an ambient temperature, $T_a$, in °C. The constants in the formula reflect a maximum junction temperature of 150°C and a thermal resistance from junction to ambient of 41.6°C/watt.

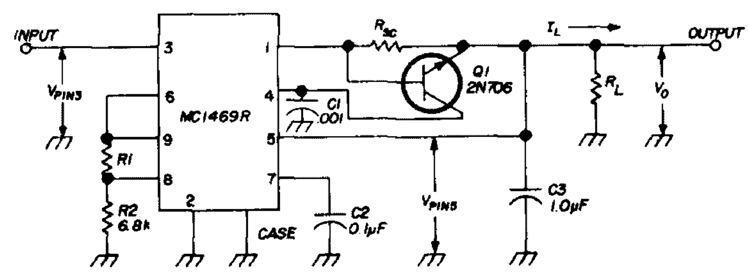



**fig. 1. Complete power-supply regulator circuit designed around the Motorola MC1469R. Components Q1 and Rsc may be eliminated if current limiting is not desired.**

## establishing boundary conditions

Now that the necessary facts have been ferreted out of the IC data sheet,[1] and the regulator circuit operation is understood, the boundaries on the volt-ampere characteristic of the unregulated input source become evident. The lower limit of the unregulated input voltage is restricted by the minimum input voltage requirement of the IC; the maximum current required from the unregulated power supply is equal to the regulator short-circuit current; the maximum voltage from the unregulated power supply must be less than the maximum IC input voltage limit, and must lie below the curve describing the ICs power-dissipation limit.

The general procedure for establishing these boundaries follows (again referring to **fig. 1**, and using the MC1469R parameters):

**1.** Specify the intended regulated output voltage, $V_o$. The value of R1 may be calculated at this time.

**2.** Specify the desired short-circuit current limit, $I_{sc}$. Calculate $R_{sc}$. $I_{sc}$ must be less than 500 mA.

**3.** Calculate $V_{pin\ 1}$ from $I_L = 0$ to $I_L = I_{sc}$.

4. Determine the minimum allowable $V_{pin\ 3}$ by adding 3.0 Vdc to the values of $V_{pin\ 1}$ calculated in step 3. If this value of $V_{pin\ 3}$ is less than 9.0 Vdc, set $V_{pin\ 3}$ equal to 9.0 Vdc.

5. Calculate the maximum allowable power dissipation, $P_{Dmax}$, at the anticipated temperature, $T_a$.

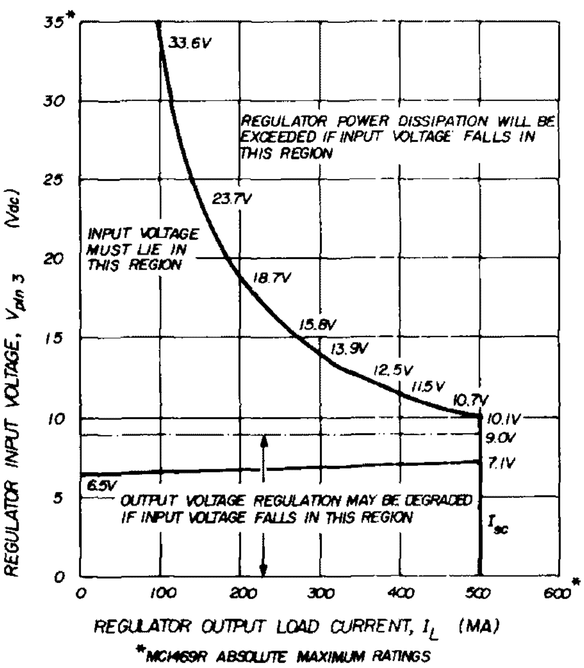



**fig. 2. Restrictions on the unregulated input voltage to the 3.5 V dc regulator circuit.**

6. Calculate the maximum allowable $V_{pin\ 3}$ for several points by substituting corresponding pairs of $V_{pin\ 1}$ and $I_L$ into the following equation

$$V_{pin\ 3} = \frac{PD_{max}}{I_L} + V_{pin\ 1}$$

Calculated values of $V_{pin\ 3}$ greater than 35 Vdc, must be set equal to 35 Vdc, since $V_{pin\ 3}$ must not exceed 35 Vdc. Note that the calculated $V_{pin\ 3}$ is somewhat optimistic, since it doesn't include the power dissipation in the IC internal control circuitry.

7. Plot the data from step 4 and step 6 on a graph of $I_L$ (from 0 to $I_{sc}$) as a function of $V_{pin\ 3}$ (from 0 to 35 Vdc). The volt-ampere characteristic of the unregulated power supply connected to pin 3 must lie within the region bounded by the two sets of data. If the two sets of data intersect, the short-circuit current specified in step 2 cannot be tolerated and must be reduced.

## design examples

Three design examples, using the step-by-step procedure previously outlined, follow.

**Example 1.** Design a regulated power supply using the MC1469R delivering 3.5 Vdc at 0 to 500 mA, operating at an ambient temperature of 25°C.

1. $V_o$ = 3.5 Vdc; R1 = $2V_o - 7 = 0$ ohms

2. $I_{sc}$ = 500 mA; $R_{sc} = 0.6/I_{sc} = 1.2$ ohms

3. $V_{pin\ 1} = V_o + I_L R_{sc} = 3.5 + 1.2 I_L = *$

4. $V_{pin\ 3}$ (min) $= V_{pin\ 1} + 3.0 = *$

5. $$P_{Dmax} = \frac{150 - T_a}{41.6} = 3.0 \text{ watts}$$

6. $V_{pin\ 3}$ (max) $= P_{Dmax}/I_L + V_{pin\ 1} = *$

* Values of these parameters for eleven different load currents, from 0.0 to 500 mA, are tabulated below.

| $I_L$ mA | Step 3 Vdc | Step 4 Vdc | Step 6 Vdc |
|---|---|---|---|
| 0 | 3.50 | (6.50)9.0 | ( ∞ )35.0 |
| 50 | 3.56 | (6.56)9.0 | (63.6)35.0 |
| 100 | 3.62 | (6.62)9.0 | 33.6 |
| 150 | 3.68 | (6.68)9.0 | 23.7 |
| 200 | 3.74 | (6.74)9.0 | 18.7 |
| 250 | 3.80 | (6.80)9.0 | 15.8 |
| 300 | 3.86 | (6.86)9.0 | 13.9 |
| 350 | 3.92 | (6.92)9.0 | 12.5 |
| 400 | 3.98 | (6.98)9.0 | 11.5 |
| 450 | 4.04 | (7.04)9.0 | 10.7 |
| 500 | 4.10 | (7.10)9.0 | 10.1 |

7. The data from steps 4 and 6 are plotted in **fig. 2**. Note that $V_{pin\ 3}$ (max) is restricted to 35.0 Vdc, and that $V_{pin\ 3}$ (min) is set equal to 9.0 Vdc. The unregulated input voltage must lie between 9.0 and 10.1 Vdc at $I_L$ = 500 mA, and between 9.0 and 35.0 Vdc when $I_L$ = 0.

**Example 2.** Design a regulated power supply using the MC1469R delivering 5.0 Vdc at 0 to 500 mA, operating in an ambient temperature of 25°C.

The calculations follow the pattern of example 1. R1 is 3.0k ohms; values for $V_{pin\ 1}$ from step 3, $V_{pin\ 3}$ (min) from step 4, and $V_{pin\ 3}$ (max) from step 6 are tabulated below. Again, $V_{pin\ 3}$ (min) must be set equal to 9.0 Vdc, and $V_{pin\ 3}$ (max) is restricted to 35.0 Vdc. The data from steps 4 and 6 are plotted in **fig. 3**.

| $I_L$ mA | Step 3 Vdc | Step 4 Vdc | Step 6 Vdc |
|---|---|---|---|
| 0 | 5.00 | (8.00)9.0 | ( ∞ )35.0 |
| 50 | 5.06 | (8.06)9.0 | (65.1)35.0 |
| 100 | 5.12 | (8.12)9.0 | (35.1)35.0 |
| 150 | 5.18 | (8.18)9.0 | 25.2 |
| 200 | 5.24 | (8.24)9.0 | 20.2 |
| 250 | 5.30 | (8.30)9.0 | 17.3 |
| 300 | 5.36 | (8.36)9.0 | 15.4 |
| 350 | 5.42 | (8.42)9.0 | 14.0 |
| 400 | 5.48 | (8.48)9.0 | 13.0 |
| 450 | 5.54 | (8.54)9.0 | 12.2 |
| 500 | 5.60 | (8.60)9.0 | 11.6 |

**Example 3.** Design a regulated power supply using the MC1469R delivering 3.5 to 5.0 Vdc (variable) at 0 to 500 mA, operating at an ambient temperature of 25 °C.

R1 is replaced with a variable resistor adjustable from 0 to 3000 ohms to allow the output voltage to be set anywhere from 3.5 to 5.0 Vdc. $R_{sc}$ remains at 1.2 ohms for current limiting at 500 mA. $V_{pin\ 3}$ (min) and $V_{pin\ 3}$ (max) from examples 1 and 2 are compared on a point-by-point basis, and the highest values of $V_{pin\ 3}$ (min) and the lowest values of $V_{pin\ 3}$ (max) become the new restrictions on $V_{pin\ 3}$, as shown *in* **fig. 4**.

## modifying the unregulated supply

One of the key points mentioned earlier was that ". . . the volt-ampere characteristic of the unregulated voltage source is the *only manipulatable system variable.*" Thus the volt-ampere characteristic of the unregulated power supply must conform to the limits established by the design requirements of the power-supply regulator. If adequate control can be exercised over the selection of the power-supply components, the unregulated power supply can be designed to conform directly to the unregulated input voltage requirements of the IC voltage regulator; if not, the unregulated power supply load profile must be modified externally.

I built an unregulated dc power supply with components on hand for use with the regulator in example 3. Its output voltage was measured at various load

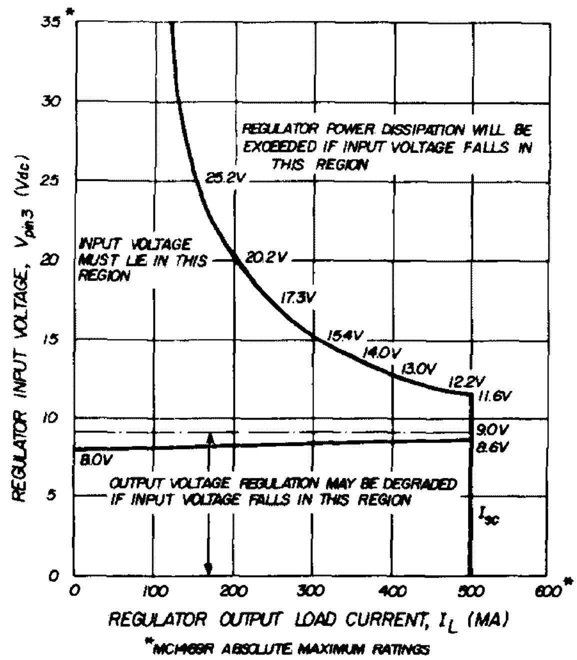



**fig. 3. Restrictions on the unregulated input voltage to the 5.0 V dc regulator circuit.**

currents and was varied from 17.9 Vdc at 0 mA to 13.9 Vdc at 500 mA. The volt-ampere characteristic of this supply is superimposed on **fig. 4** and identified as the "unacceptable input voltage profile." The output voltage of this supply would have exceeded the maximum allowable voltage on pin 3 of the MC1469R, under the specified design conditions, and the IC power dissipation would have been exceeded when the regulator output-current demand increased much beyond 270 mA. A substantial portion of the total output current capability of the voltage regulator IC could not have been used.

A 10-ohm resistor placed in series with

the positive lead from the unregulated power supply modified the power-supply load profile, producing the "typical input voltage profile" volt-ampere curve in fig. 4. Note that the voltage applied to pin 3 of the MC1469R remains in the permissible input voltage region for all values of $I_L$, allowing the full capability of the voltage regulator IC to be realized.

Experimental measurements on this voltage regulator system revealed an MC1469R junction temperature rise of 86 °C at $V_o$ = 3.5 Vdc and $I_L$ = 500 mA (2.8 watts dissipation). The measured junction temperature rise was 30 °C below the calculated value; this is attributed to (a) the conservative thermal resistance values used in the IC data sheet, and (b) the additional thermal path created by attaching the IC case to the mounting studs of the particular socket I used.

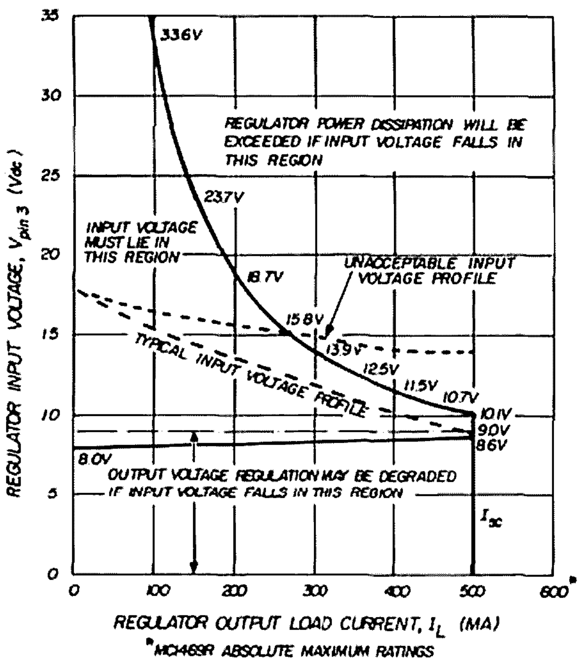



**fig. 4. Input voltage restrictions and experimentally-determined unregulated power supply volt-ampere characteristics for a 3.5 to 5.0 V dc variable regulated power supply.**

## reference


1. "Motorola MC1469R Positive-Power-Supply Voltage Regulator Integrated Circuit Data Sheet," Motorola Semiconductor Products, Inc., P. O. Box 20912, Phoenix, Arizona 85036.


ham radio

# lunar-path nomograph

An aid for determining signal attenuation due to variance in earth-moon distance

Bill Ress, WA6NCT, Eimac Division of Varian, San Carlos, California 94070

Here's a handy operating aid that will be appreciated by the moonbounce gang. It allows accurate determination of the variance in free-space signal loss caused by the change in earth-moon path length.

The distance between earth and moon at the moon's perigee, which is the point of the moon's orbit closest to earth, is nominally 221,463 miles. At apogee, or the point of the moon's orbit farthest from earth, the nominal distance is 252,710 miles. This variation can account for a 1.28-dB change in signal level for the earth-moon path, or a total of 2.56 dB for the round trip.

The optimum time for moonbounce work, of course, is at the moon's perigee. However, since the earth-moon path distance varies, it's convenient to know the expected signal levels at times other than perigee.

## data forecasts

It will be necessary to know the distance between earth and moon at the time of interest. Various tables can be consulted, but the most convenient source of this data is *Sky and Telescope*,[1] a magazine published for astronomy enthusiasts. In the Celestical Calendar section of this magazine, perigee and apogee times of the moon's orbit are given, together with earth-moon distances and the moon's effective diameter. This data is published a month in advance, since perturbations in the moon's orbit cause perigee and apogee to vary from month-to-month. Other useful information is also given. Sunspots and times of meteor showers are forecast, and articles are published on radio astronomy and satellite activity.

## using the nomograph

**Table 1** is a computer printout giving free-space signal attenuation in dB for

table 1. Computer printout showing free-space path attenuation for several frequencies as a function of various earth-moon distances.

| distance (miles) | 50 MHz (dB) | 144 MHz (dB) | 432 MHz (dB) | 1296 MHz (dB) | 2375 MHz (dB) | 3600 MHz (dB) |
|---|---|---|---|---|---|---|
| 220000 | 177.82 | 187.01 | 196.55 | 206.10 | 211.36 | 214.97 |
| 225000 | 178.02 | 187.21 | 196.75 | 206.29 | 211.55 | 215.16 |
| 230000 | 178.21 | 187.40 | 196.94 | 206.48 | 211.74 | 215.36 |
| 235000 | 178.40 | 187.58 | 197.13 | 206.67 | 211.93 | 215.54 |
| 240000 | 178.58 | 187.77 | 197.31 | 206.85 | 212.11 | 215.73 |
| 245000 | 178.76 | 187.95 | 197.49 | 207.03 | 212.29 | 215.90 |
| 250000 | 178.93 | 188.12 | 197.66 | 207.21 | 212.47 | 216.08 |
| 255000 | 179.11 | 188.29 | 197.84 | 207.38 | 212.64 | 216.25 |

several frequencies and earth-moon distances. The attenuation numbers can be read directly from the table for earth-moon distances between 220,000 and 255,000 miles in the 5000-mile increments. The nomograph (**fig. 1**) is used for linear interpolation to determine the free-space signal attenuation for earth-moon distances not given in the table. Note that the table and nomograph are based on an earth-moon distance of 220,000 miles.

frequency, then add this number to the interpolation factor. Example:

Earth-moon distance = 233,000 miles

Operating frequency = 2375 MHz

From **table 1** the free-space loss (220,000 miles) = 211.36 dB

From **fig. 1** the interpolation factor (233,000 miles) = 0.50 dB

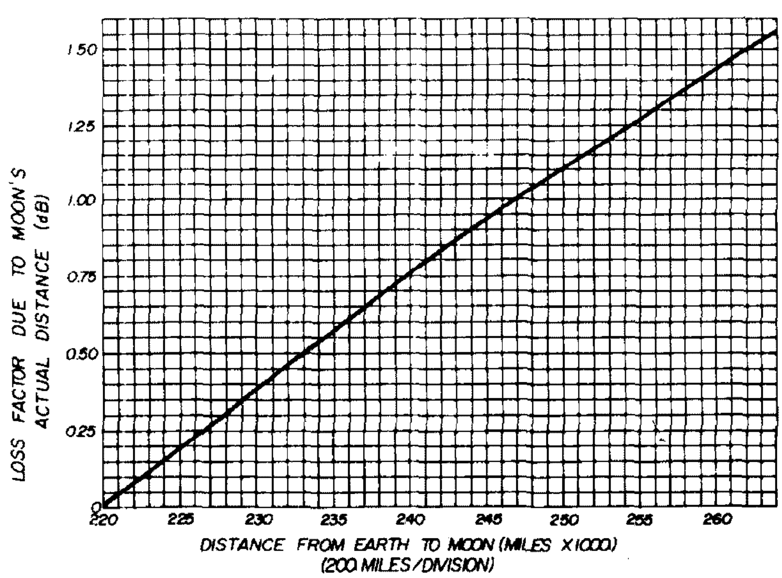



**fig. 1. Nomograph for determining loss (interpolation) factor for actual distances between earth and moon.**

## example

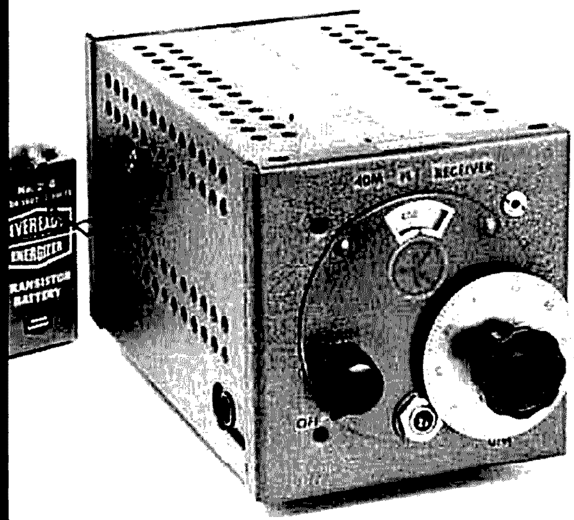
Knowing the earth-moon distance, enter **fig. 1** at the distance, and move vertically to the intercept line. Then read the interpolation factor on the ordinate. From **table 1**, determine the free-space loss at 220,000 miles for your operating

Therefore, the free-space attenuation for 233,000 miles = 211.36 + 0.50 dB = 211.86 dB.

### reference


1. *Sky and Telescope,* Sky Publishing Corp., 49-50-51 Bay State Road, Cambridge, Massachusetts 02138.


**ham radio**



# converted BC-1206 for 7 MHz

Another surplus item easily modified with transistors for low-frequency work

Bill Latta, Jr., W4FIN, 3822 North Western Parkway, Louisville, Kentucky 40212

The BC-1206 beacon receiver is another of the fine units still available on the surplus market that lends itself to modification for amateur use. It tunes from 200 to 400 kHz and has a 135-kHz i-f. The original design was for a 28 Vdc power supply.[1,2]

For some time I had been thinking about adapting a BC-1206 for use as a portable ham receiver. It looked like a natural for solid-state conversion; so I replaced the tubes with transistors, added a solid-state bfo, 40-meter crystal converter, and a manufactured audio stage. Result: an all-solid, completely portable receiver.

## construction

Preliminary work consisted of checking the unmodified receiver on 28 Vdc, then on 12 Vdc after paralleling tube filaments. Reception of local and distant stations was good. It would have been possible to use the receiver on 12 volts with only a converter and bfo, but I chose the solid-state route for the challenge and because of economic reasons.

## beat frequency oscillator

The bfo (fig. 1) is a miniaturized version for 135 kHz adapted from reference 3. It's mounted in a sawed-off TV i-f can, to which solder lugs were attached and bent outward for soldering to the inside front panel. I wired the components around the original i-f coil form after removing the wire. The subminiature rf choke was mounted inside the form by drilling a small hole through the form and pulling one choke lead through the hole. The original slug can be used to

control pitch. If no pitch control is desired, the slug can be removed.

I removed the original phone jack and substituted a ¼-inch panel bushing. The bfo assembly was mounted with the pitch control protruding through the bushing for front-panel access.

The bfo is well shielded, and a pi-network filter is included in its power lead to minimize stray radiation. I used miniature coax cable for the output lead. In the completed receiver, a "birdie" was picked up at the bfo's second and third harmonics. A phase-shift oscillator, which would have no harmonic output, should alleviate this problem and is a possible improvement. The receiver is used for cw and ssb only, so I didn't include a bfo control switch.

## the converter

The 40-meter converter (fig. 2) is a miniaturized one-band version of another design.[4] It is housed in a steel box that fits into the space left by the i-f detector, the audio tubes and the output transformer.

I drilled a hole in the side of the crystal socket and mounted the socket horizontally to provide clearance for the housing. No special precautions are necessary in construction except to keep L1 and L2 axes at right angles to minimize coupling.

Since my interest was in the 7- to 7.1-MHz range, and receiver bandwidth is best between 200 and 300 (on the main tuning dial), a 6.8-MHz crystal was used. The rf amplifier isn't needed for sensitivity, but it helps to suppress images.

## detector and audio

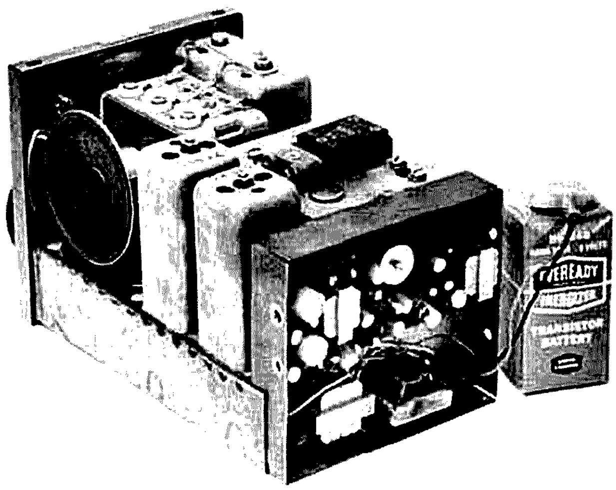
A single 1N270 could be used as a detector, with the bfo voltage injected at the i-f amplifier gate. However, I decided to use a product detector.[5] Experimentally, a pnp germanium transistor was a good substitute for the two diodes (fig. 1).

I used a Round Hill model AA-100 audio amplifier* because I had one on hand. However, other audio boards, modules, or ICs will work just as well.

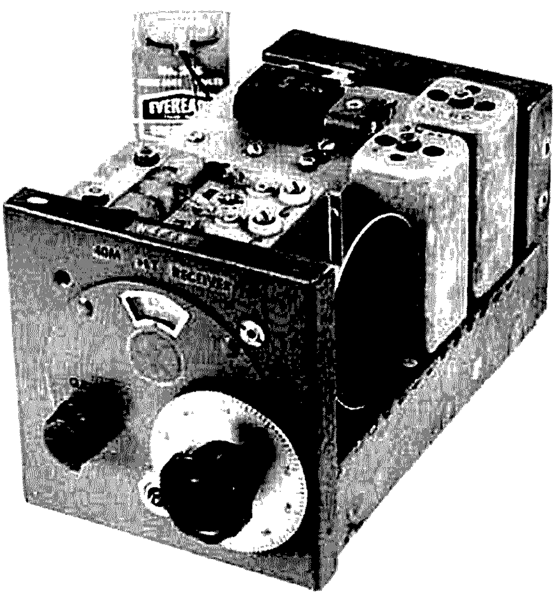
I mounted the audio amplifier on the rear deck of the receiver after removing all the original rear components except the small open-frame filter choke. To make the amplifier fit into the space, it was necessary to remove the output transformer, relocate one 100 $\mu$F capacitor, then saw off both ends of the board.

*Round Hill Associates, Inc., 434 Avenue of the Americas, New York, N. Y. 10011.

**Rear view showing how the audio deck is mounted. Battery is an Eveready 246; six size D cells in series could be used.**



**Interior view of modified receiver. Converter is in box with crystal on top.**



I put a thin sheet of insulation on both sides of the board and cemented more insulation to the inside back of the amplifier housing to ensure against short circuits.

and rf stages will tend to oscillate if jfets are used. Rather than neutralize these circuits, I used 3N128 mosfets, which provide good stability and adequate gain. The usual precautions must be used when

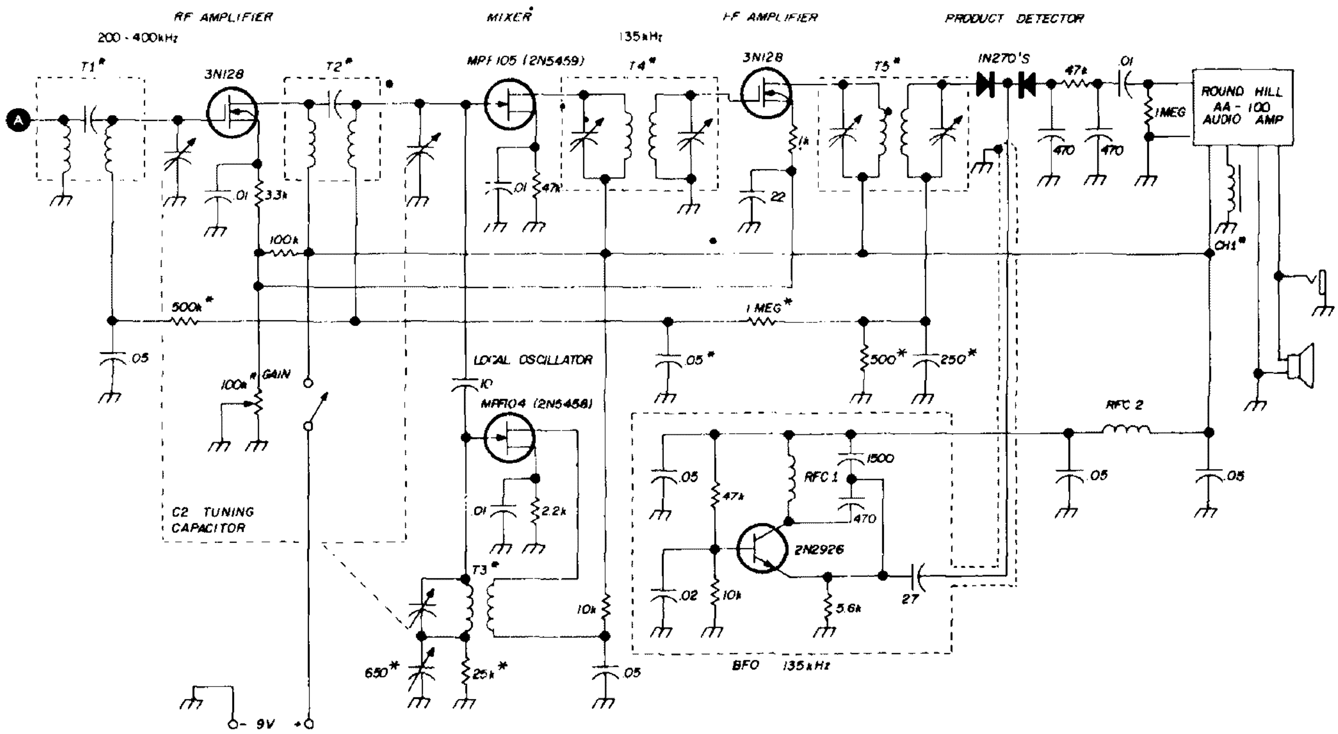



**fig. 1. Schematic of converted BC-1206 beacon receiver with added bfo and audio stages. Components marked with an asterisk are original. RFC1 and RFC2 are 3.9 mH (J. W. Miller 70F393AI).**

Next, I solder-mounted the output transformer below the board and reconnected the transformer. Since the board operates from positive ground, the input and output windings must be lifted from board ground and connected to the main radio chassis ground to avoid shorting the battery. These are pins 1 and 9 respectively of T1 and T3 on the AA-100 board.

The amplifier oscillated when it was connected to the common battery source, and the usual decoupling methods didn't help. The problem was cured by connecting the open-frame filter choke in series with the amplifier's negative lead. A 2¼-inch 3-ohm speaker was mounted in the space left by the original rf and converter tubes.

## i-f and rf amplifiers

The receiver will work merely by replacing tubes with fets; however, the i-f handling such devices. I used an MPF105 (2N5459) and an MPF104 (2N5458) for mixer and local oscillator.

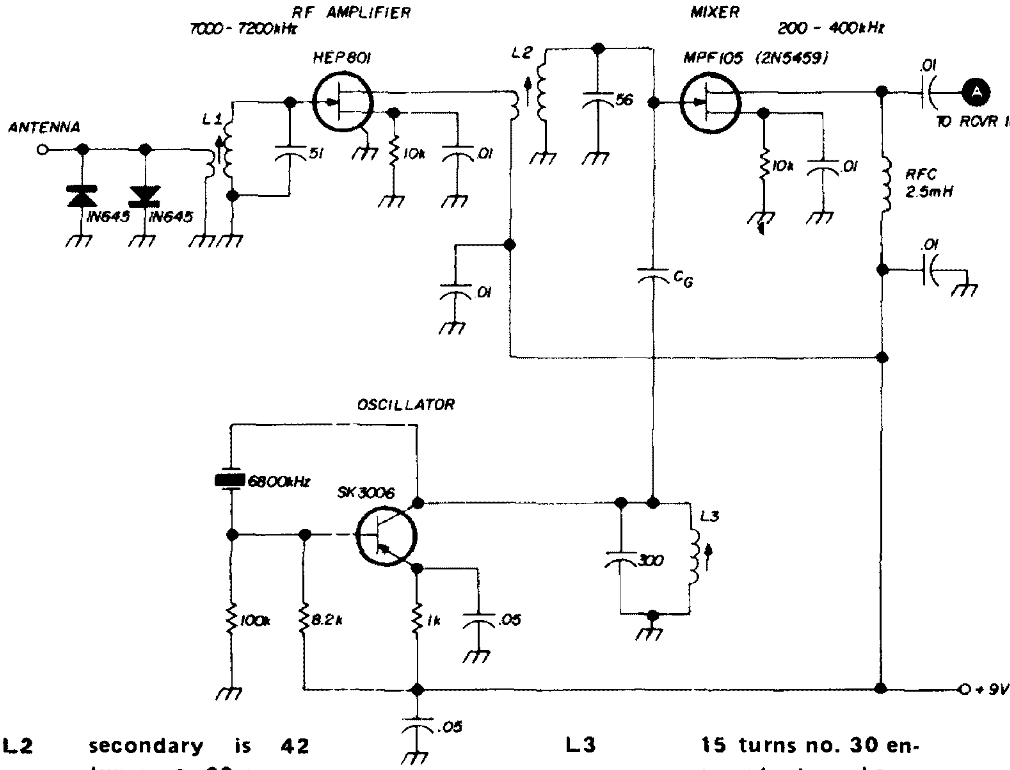
I varied resistances for each stage for optimum performance, then soldered fixed resistors into the circuit. The agc line was left intact, although it has little effect on operation. Better agc might be had by using double-gate mosfets.

## power supply

Optimum supply voltage is 9 volts, although the receiver will operate on 7½ volts with less audio. With 12 volts, additional spurious signals will appear in the output. These are caused by the crystal oscillator beating with the local oscillator harmonics. At 9 volts, idle current is about 15 mA with peak current over 35 mA.

Small transistor batteries will work, but they have poor regulation and short

life for this application. The larger transistor-radio batteries are better. Nicad or mercury batteries are also good. Six size D flashlight cells in series make an economical power pack.




L1, L2 secondary is 42 turns no. 30 enameled, closewound on ¼" slug-tuned form; primary is 7 turns no. 30 enameled, closewound over cold end of secondary.

L3 15 turns no. 30 enameled, closewound on ¼" slug-tuned form

**fig. 2. Solid-state converter for 7 MHz. Components fit into space formerly occupied by tubes and output transformer. "Gimmick" is two 1" lengths of hookup wire twisted together.**

## the exterior

I drilled additional holes in the right side of the case for a speaker grill and added a hole in the left side for the antenna connection. I also added bumper feet to the bottom, a cutout in the rear to pass the battery connector to the outside, and mounted a battery holding strap onto the rear of the cabinet.

## operation

The converted BC-1206 provides sensitive, stable, and fairly selective cw and ssb reception in a small package with low power requirements. It needs only a short antenna: a wire about 15 feet long or so.

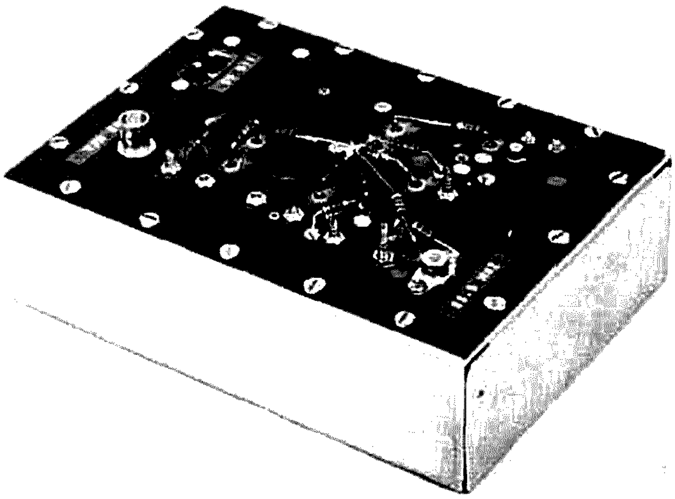
I've used this receiver with a companion low-power transistor transmitter with good results, which satisfied an urge for a complete, miniaturized, all-battery-operated, solid-state station.

Low-power transmitters are fun, but it's more fun using a low-power receiver with the transmitter. The receiver is also a valuable adjunct to your station for emergency use.

### references

1. Grayson, "Surplus Schematics Handbook," Cowan Publishing Corp., Fifth Printing, August, 1967, p. 49.
2. "Instruction Manual, Model 524 Beacon Receiver," Setchell Carlson, Inc., St. Paul, Minnesota. (Sometimes available with surplus receivers.)
3. Marriner & Meredith, "Transistorized 455-kHz BFO," *ham radio*, July, 1968, p. 12.
4. Sam Creason, "An FET Converter for 40 and 160," *73*, October, 1967, p. 26.
5. "Heterodyne and Product Detectors," The Radio Amateur's Handbook, ARRL Staff, 1969 Edition, p. 97.

ham radio



# low-noise converter for 432 MHz

A circuit using inexpensive fet's that gives a good account of itself over the long hauls

James W. Brannin, K6JC, 424 Anson Avenue, Rohnert Park, California

Now that economically priced solid-state devices are available for use above 400 MHz, the cost of building uhf equipment has been considerably reduced. The converter described here uses only seven active devices, all of which cost less than $3.50. The 2N5245s net at 70¢, the 40237s are 38¢, and the diodes are about 30¢ apiece.

The bandwidth of the converter at the 3-dB points is between 431 and 433 MHz. The noise figure is 3.5 dB. If you'd like to reduce this another half dB or so, a 2N5397 fet can be used in the first rf stage instead of the 2N5245; however, the 2N5397 costs about $8.50.

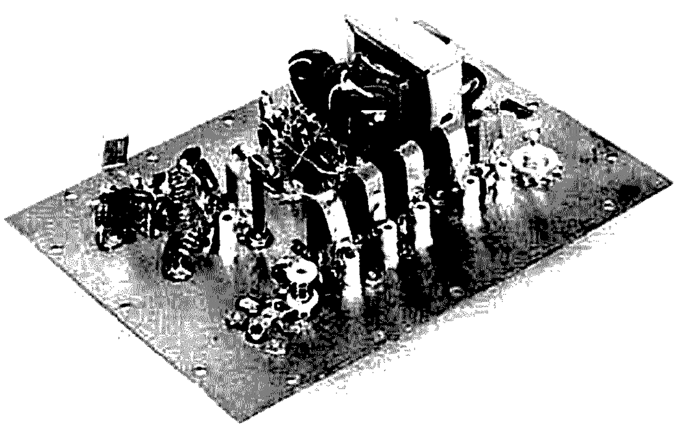
The circuit is conventional (fig. 1). Two fet grounded-gate amplifiers and a common-emitter mixer are used. Oscillator output is about 8 mW. The quadrupler output is somewhat below this, but it provides sufficient injection voltage for the mixer. Both input and local oscillator signals are applied to the mixer transistor base. This arrangement offers the best conversion efficiency for this type transistor at these frequencies.

## layout and construction

Component arrangement, component spacing, and wiring placement have been optimized for best performance based on several earlier versions. Shielding isn't necessary, since there appears to be enough feedback in the front end to eliminate any tendency toward instability. As a matter of fact, the shielding that was installed around each of the first three stages in the original model had to be removed to eliminate regeneration. The spacing between tuned circuits in the amplifier and mixer was optimized to provide satisfactory interstage coupling without the use of capacitors.

The chassis measures 5x7x2 inches. The circuit board is glass epoxy with copper foil on one side. These boards are available from many surplus outlets. Some are pullouts from computer modules and have discrete components mounted on them, which can be salvaged for your junk-box inventory.

**Component arrangement. Rf amplifiers and mixer tuned circuits are mounted along the near edge of board; oscillator and power supply are at rear.**



All wiring except that for the power supply is on top of the board. This allows power to be fed to all stages via feedthrough *capacitors*, which minimizes coupling problems in the wiring. The board should be drilled as closely as possible to the dimensions on the template except for holes associated with the power supply.

## coil construction

The Q of the tuned circuits was increased when the hairpin loops were made of copper strip instead of wire. Coils L1, L2, L4, and L9 are made by cutting 20 gauge flashing copper in strips ¼ inch wide by 2-1/8 inches long. The mixer coil, L5, should be ¼ inch shorter to reduce its inductance so that it will resonate with the tuning capacitor piston halfway inserted. The end of the coil that is soldered to the feedthrough capacitor is trimmed to a point. When the loops are installed they should stand 1-3/8 inches above the board.

Coil L3 is made from a copper strip 2¼ inches long. The ground end should not be trimmed to a point but should be bent out 1/8 inch for soldering to the board. Otherwise its construction is the same as the other amplifier coils.

The i-f output coil, L6, consists of 18 turns of no. 24 enamelled wire close wound on any type of 3/16-inch diameter slug-tuned form. (A grid-dip oscillator should be used to check the coil.) The output link, L7, is made from three turns of no. 22 enamelled wire wound on the cold end of L6.

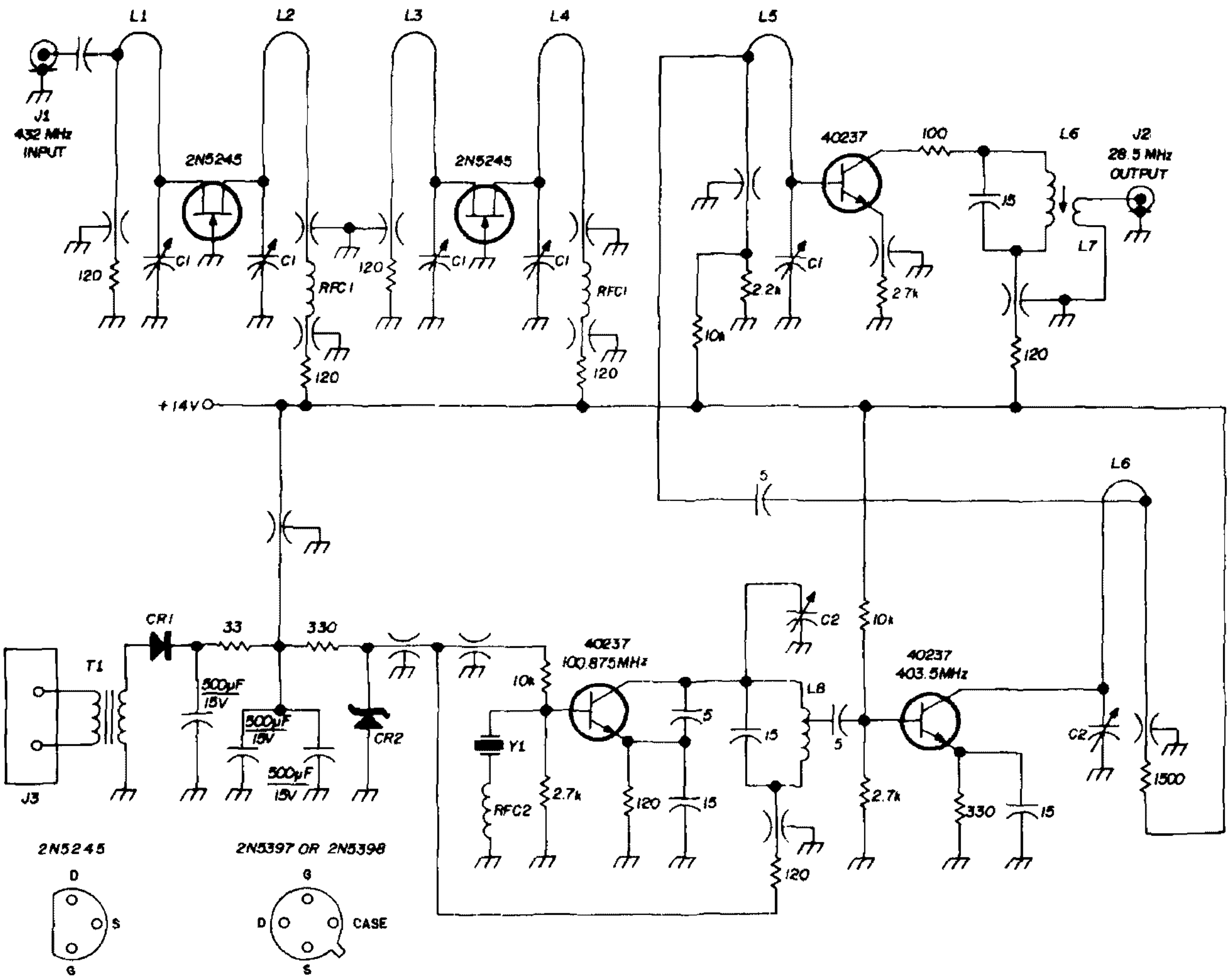
The oscillator coil, L8, consists of 4¾ turns of no. 20 enamelled wire, space-wound, ½ inch long by ¼ inch I.D. The output tap is located 1 turn from the top.

## local oscillator

An rf choke is used in the crystal circuit to allow warping of the crystal frequency without appreciable loss in output. I suggest the crystal be ordered for 100.872 instead of 100.875 MHz, as the frequency will fall much closer to 100.875 if the crystal is ground for this lower frequency. The oscillator operates from the 10-V regulated supply, and the stability after a few minutes warmup is excellent. The frequency at 432 MHz was checked and exhibited a drift of about 300 Hz over a 3-hour period.

## assembly

The unit should be assembled in the following sequence after the board is drilled.




| | |
|---|---|
| C1 | 1-8 pF piston variable (Triko 106-01M) |
| C2 | 1½—10 pF air padder (Johnson T106S) |
| CR1 | any diode rated at 200 V piv |
| CR2 | 10 V zener |
| J1 | UG-1094A/U BNC receptacle |
| J2 | phono jack |
| J3 | standard ac receptacle to accommodate TV cheater cords |
| RFC1 | 7-3/8" length no. 24 wound on ½-watt resistor |
| RFC2 | Same as RFC1 (may not be necessary) |
| T1 | 12.6 V filament transformer (Calectro D1-750 or equivalent) |
| Y1 | *100.872 MHz 5th overtone crystal (see text)* |

fets are Texas Instruments 2N5245 or Siliconix 2N5397

transistors are RCA 40237

feedthrough caps are Allen Bradley FW5N, 470 pF

standoffs are Arco DM15 1501

chassis is California A101, 5 x 7 x 2 inches

fig. 1. Schematic of the 432-MHz solid-state converter. Circuit features low-noise front end and highly stable oscillator.

1. Install all feedthrough capacitors.

2. Install the transistor sockets, using a good grade of cement.

3. Install the piston tuning capacitors and connect them to the proper terminal on the sockets.

4. Install the output slug-tuned coil.

5. Install the oscillator coil. This coil is rigidly supported by mounting it in a vertical position. Solder the bottom end to the center of the feedthrough capacitor, and connect the top end to C2 and the collector pin on the transistor socket using short, stiff leads.

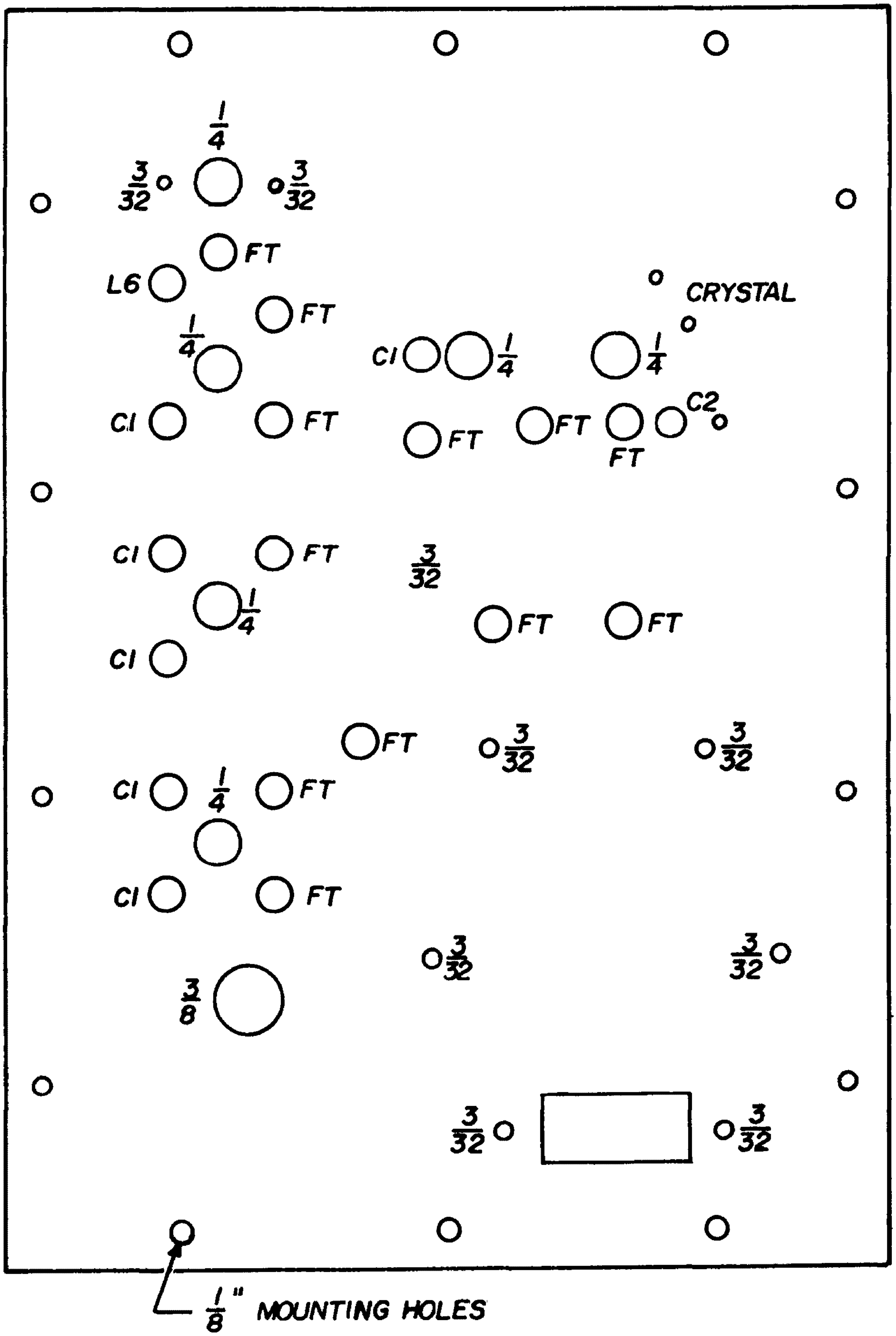



fig. 2. Full-size drill template for the 432-MHz converter. Holes for feedthrough capacitors (FT), L6 and C1 are 3/16" diameter. BNC connector holes are 3/8". Transistor sockets and J2 require 1/4" holes.

6. Install the hairpin loops, being careful to solder them to the feed-through capacitors so that the capacitor isn't shorted with a drop of solder. The coupling capacitors on L1, L5 and L9 are soldered onto the loops near the top on the bypassed side.

7. Complete the quadrupler and oscillator wiring.

8. Finally, install the power supply, then complete the wiring on top of the chassis.

The two fets should be installed with the input tuned circuit connected to the source and the output connected to the drain so that 2N5397 or 2N5398 fets can be used if desired. When the 2N5245s are installed, the drain and source will be reversed; however this will have no effect on their operation. This is not true with some of the other fets.

## tuneup and adjustment

If a commercial-type signal generator isn't available, I'd suggest that one of the vhf small-signal sources be built in accordance with my article in a previous issue of *ham radio.*[1] The converter can be aligned using the third harmonic from a 144-MHz transmitter. However, it may be difficult to obtain optimum performance unless a variable output generator of some type is used.

The first testing should begin with the oscillator and quadrupler. A gdo tuned to the oscillator frequency will allow you to determine if the crystal is operating on the proper mode. A vtvm with an rf probe can be used to peak the quadrupler output. The probe should be coupled through a 33-pF capacitor to the top of the mixer hairpin loop and the quadrupler adjusted for maximum output. It will be necessary to use one of the lower-voltage scales to detect the very small amount of rf at this point in the circuit. Once the signal from the generator or from the 144-MHz transmitter is found, all tuned circuits can be readjusted for optimum signal.

A word of caution here: When peaking the circuits, tune off the incoming signal and check the noise level as indicated on a receiver S meter. The objective is to get the greatest possible signal-to-noise ratio. Final adjustments can be made with a noise generator, which should be used if possible. A second word of caution: Use a noise generator for *very slight readjustments only.* You may get one of the stages tuned to the image frequency and the noise figure will appear to be very good, but 432-MHz stations probably won't be heard unless they're next door.

The final adjustment should be made with the converter mounted in the chassis with all of the mounting screws tightened carefully. Final adjustments can sometimes be made quite well while listening to a distant weak signal provided it isn't fading.

## operating results

Several of these converters have been operating in this general area with very good results. Noise figure has been measured at close to 3.5 dB on all units. I have never failed to copy WB6PDN's two watter in Stockton with this converter connected to a 16-element colinear antenna only 24 feet high. Stockton is well over 100 miles from my location, and the path is across several mountain ranges. Stations in the San Francisco bay area as far south as San Jose, a distance of about 80 miles, are worked regularly. Only about four stations use more than a few watts on 432 MHz.

While converters in this band are not as easy to build and adjust as those for 144 and 50 MHz, they will work without too many difficulties if one is careful in construction. Try to use a hefty signal when starting the alignment and graduate later to weaker signals.

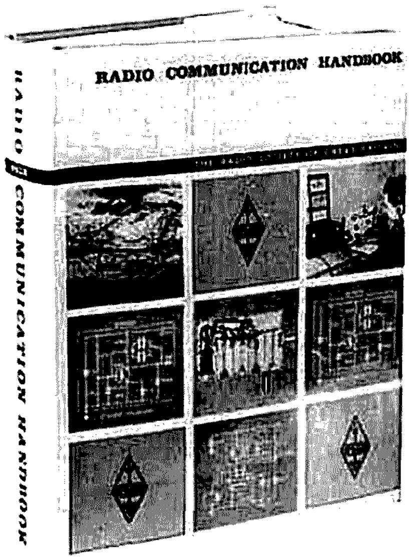
My thanks to many of the dedicated vhfers in this area who provided much helpful information in building this converter.

### reference

1. James W. Brannin, K6JC, "A Stable Small-Signal Source for 432 MHz," *ham radio,* March, 1970, p. 58.

**ham radio**

# partners in excellence



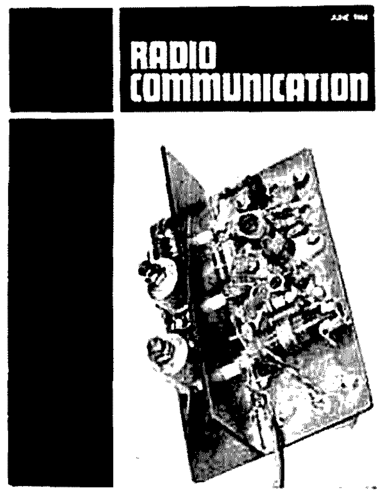

An outstanding technical guide to all phases of amateur radio. In 832 pages 20 complete chapters are devoted to such subjects as single sideband, antennas, mobile equipment, RTTY and much, much more.

This excellent book has received wide acclaim on both sides of the Atlantic and belongs in your library . . . **now.** **$11.95**

**OTHER**
**POPULAR RSGB PUBLICATIONS**

| | |
|---|---|
| Radio Date Reference Book | $2.50 |
| VHF-UHF Manual | 3.75 |
| Amateur Radio Techniques | 2.50 |
| Amateur Radio Circuits Book | 2.00 |
| World at Their Fingertips | 2.50 |

All prices postpaid in USA & Canada




Many thousands of you have become very familiar with the various Radio Society of Great Britain books and handbooks, but very few of you are familiar with their excellent magazine, **Radio Communication.**

This is the oldest and most widely read British amateur radio magazine. Published monthly it provides complete coverage including such popular features as: Technical Topics, a monthly survey of the latest ideas and circuits, Four Meters and Down, a rundown of the latest in VHF and UHF and much more.

It includes numerous technical and construction articles in addition to a complete rundown on the month's events in amateur radio. Surely a most interesting addition to your amateur radio activities.

**Act Quickly!** A price increase will be announced very soon. Send your subscription to us today while it is still only **$6.00** per year.

## book division
## comtec

Box 592 • Amherst, New Hampshire 03031

**"WE ARE THE EXCLUSIVE NORTH AMERICAN DISTRIBUTOR FOR RSGB PUBLICATIONS — DEALER INQUIRIES INVITED"**

More Details? CHECK—OFF Page 94

# precise frequency tuning with ssb equipment

Hints on pinpointing operating frequency in the ssb, cw, and rtty modes

Doug Horner, WØKD, 1260 E. Avenue, Marion, Iowa 52302

Single sideband has come of age. That struggling infant, the "new" communications mode that initially attracted only a select group of amateurs, has finally been accepted. Gone are the somewhat snobbish (but enjoyable) sideband dinners and sideband forums that used to be so popular. Articles explaining single sideband are now conspicuous by their absence in the amateur literature.

Despite the wide acceptance of ssb, some of its aspects remain mysterious and confusing to many. One is how to determine the *actual* transmitted and received frequency, a capability especially needed by net-control stations and to a lesser degree by net-responding stations.

This article reviews some of the basic principles unique to ssb and offers suggestions to aid in adjusting transmitters and receivers to a desired frequency with certainty and precision.

## background

Sideband started on the ham bands with home-constructed exciters. Then we added linear amplifiers, vox, bandswitching, and the oscilloscope to check linearity. Receivers became more complex. We added sideband slicers, compression amplifiers, Q multipliers, etc., until a sidebander's station looked like the bargain basement in a radio store.

Then manufacturers made exciters available--receivers too. These were followed by linears. In 1957 Collins introduced a revolutionary little set that became the vanguard of the industry—a transceiver. The KWM-1 combined both transmitter and receiver in one compact package, and the radio-store look of the ham shack gave way to a neat, wife-approved station.

Sideband transceivers so dominate ham radio today that even seasoned hams at times feel as though they are operating a-m equipment in unfamiliar territory. The following paragraphs will help to

dispel some of the confusion associated with tuning ssb equipment and allow you to determine operating frequency whether using voice, cw, or rtty.

## a-m signals

For a start, let's look at the grandpa signal, amplitude modulation. Basically, any radiotelephone transmitter merely translates audio frequencies to radio frequencies, and the receiver translates the radio frequencies back to audio frequencies. To accomplish this, one must use a frequency translator into which is injected two signals: that which is to be translated and another signal generated locally In an oscillator. When combined in the frequency translator, the two signals are said to be heterodyned. Sum and difference signals are produced, and the composite signal consists of the cw carrier and two mirror-image sidebands, which contain the a-m signal. A graphical representation of this is shown in fig. 1.

## ssb

Those who take license exams must learn that, "For each modulating frequency an upper and lower frequency appears either side of the carrier by an

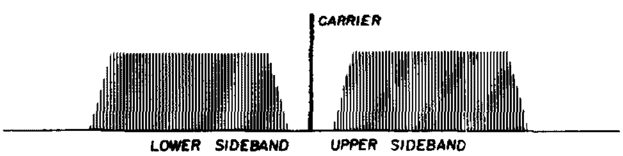



**fig. 1. Representation of amplitude-modulated signal, consisting of carrier and two sidebands.**

amount equal to the frequency of the modulating tone." For speech, a band of frequencies (roughly from 300–3000 Hz) appears rather than individual side frequencies. These are called sidebands.

As the magazine articles told us when ssb was new, "No information is passed by the carrier. All it does is 'blow the whistle.' Therefore, we don't need the carrier for communications. Also, since it contains all the information needed for communications." This concept is depicted in fig. 2. The upper sideband remains, as in the a-m signal, but the carrier and lower sideband have been suppressed.

## ssb tone generator

A statement similar to that made earlier about an a-m signal can be made about an ssb signal: "for or modulating frequency an upper and lower sideband appears, which is removed from the suppressed carrier by an amount equal to the frequency of the modulating tone."

The practical result is that, if you feed a pure tone of a given audio frequency into a quality ssb transmitter, a single

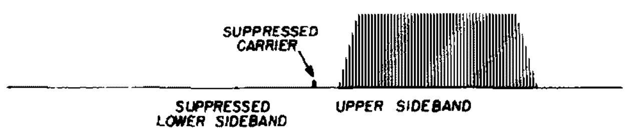



**fig. 2. Single-sideband signal; one sideband and the carrier have been suppressed.**

frequency will be produced that is removed from the suppressed carrier by an amount equal to the audio frequency. Many ssb transmitters and transceivers take advantage of this phenomenon by using a tone generator to derive a cw signal.

Now here's where the confusion begins. If you calibrate your receiver so the dial hairline coincides with 0 on the dial when you tune to zero beat against the calibrator signal, what you actually hear is the suppressed-carrier frequency. Theoretically, no energy is either radiated or received on that frequency. If you're operating phone on upper sideband, communication is taking place 300–3000 Hz above the dial reading. If on lower sideband, the energy is 300–3000 Hz below the dial frequency.

## cw frequency spotting

The explanation above is fairly easy to understand, but let's say you own a transmitter that generates cw by passing a

tone through the balanced modulator and the sideband filter. You're the control station for a cw net, say, and must place your transmitter on a specific frequency to begin calling the net. How do you set your dial to obtain that frequency?

This isn't too difficult. If you know the frequency of the tone generator, and the upper-sideband position is used for cw, then tune your dial lower in frequency by an amount equal to the frequency of the tone oscillator (see fig. 3). If the audio tone is converted to "carrier" and the true carrier is suppressed, to be on frequency you must sidestep the dial tuning to place the resultant "carrier" where the true carrier would have been if it were radiated.

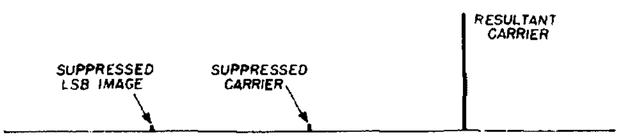



**fig. 3. Frequency setting on cw with ssb transmitters using a tone generator. In the example shown, usb is used; dial setting must be set lower in frequency by an amount equal to tone oscillator frequency.**

As an example, you're to call a cw net on 3565 kHz, and your sideband transmitter generates cw with a 1500-Hz tone oscillator working against the upper sideband bfo crystal. By setting your dial to 3563.5 kHz, your output frequency will be 3565 kHz;

| | |
|---|---|
| suppressed carrier frequency | = 3563.5 |
| tone oscillator frequency | = + 1.500 |
| | 3565.000 |

## frequency setting on rtty

Now let's suppose you're a control station for a tty net and you're feeding tones into your sideband transmitter, which it converts to frequency-shifted rf for transmission. How do you set the transmitter on frequency?

Usually, but not always, a center frequency straddled by the two tty frequencies is assigned as net frequency. It's possible your ssb exciter won't pass audio frequencies above about 2500 Hz, so you use a tone generator that operates on 1275 and 2125 Hz. This gives the more commonly used 850-Hz shift. One-half of 850 Hz is 425 Hz; 2125 minus 425 is 1700, 1275 plus 425 is 1700. Therefore, from this we may conclude 1700 Hz is the center frequency. To straddle the assigned frequency, first zero-beat your dial against the nearest 100-kHz point, then tune to 1700 Hz above the assigned frequency with the exciter and receiver in the lsb position. With 2125 and 2975 tones, the above information applies, except the center frequency is now 2550 Hz. Therefore, you should tune 2550 Hz higher for these tones.

Most sideband exciters neither have dials that can be read this accurately, nor does the output frequency ordinarily coincide this closely with the dial reading; but this is good ballpark information for much of the equipment available today. Very exacting requirements will demand the use of a counter or other accurate frequency meter to set up the individual transmitted frequencies precisely.

## mark and space signals

Reference has been made to operating rtty on lower sideband. This is an arbitrary practice followed by amateurs, MARS, and some military organizations on the hf bands. What actually is transmitted can be remembered easily by stealing an acronym from a tobacco company, "LS/MFT." Supply the words, "low space makes fine teletype." The mark frequency is always the higher frequency transmitted, and space is the lower frequency. In practice the space signal sounds higher, but that's because of sideband inversion due to operating the equipment on lower sideband. When a transmitter employing a carrier is used, such as the a-m and fm equipment on the vhf bands, the tones are broadcast "right-side-up" with the space tone higher. This is known as afsk, or audio frequency-shift keying.

**ham radio**

# the simplest audio filter

These circuits offer the ultimate in simplicity for effective cw reception

E. Dusina, W4NVK, 571 Orange Avenue West, Melbourne, Florida 32901

Many audio filters have appeared in magazine articles for both cw and ssb. The more popular of these have been centered around 44- and 88-mH toroids available for as little as 30 cents each from surplus sources.* This article describes such a filter for single-tone cw reception, but with all the nonessential frills removed. Also these filters have been tested on an audio-frequency spectrum *analyzer*, so their response using actual speaker loads is known quite exactly.

## filter criteria

A cw filter must have one prime qualification to be considered good, namely very narrow bandwidth. Since 20 wpm represents only about 10 Hz actual information bandwidth, the filters of use to hams can be very sharp indeed. However, we can't practically use a 10 Hz filter, because transmitter plus receiver drift would require constant retuning. Also the filter would ring, producing disagreeable-sounding code. The ringing could be removed by damping, but the pulses would still have rounded, sine-wave-like shapes. Therefore, the filter should be somewhat wider than the minimum bandwidth based on information theory.

Assuming ssb-grade frequency stability, the combined short-term drift of transmitter and receiver will seldom exceed 50 Hz. This bandwidth, being five times wider than the minimum necessary to pass all the information, will pass pulses with a square-wave shape. So a practical cw filter should not be narrower than about 50 Hz; and to be called good, it should not be wider than about 100 Hz.

Another desirable feature for a general-purpose filter is that it operate in the speaker leads to eliminate the necessity for modifying resellable hardware.

If you're using an ssb receiver, all of which have linear product detectors, there is absolutely no signal-to-noise ad-

*See flea market ads in practically any issue of *ham radio.*

vantage whether the filter is in the predetection (i-f) or postdetection (audio) circuits. This isn't true for a-m receivers with their conventional detectors, however. For these, a Q multiplier in an i-f stage will give the ultimate in weak signal reception. But the audio filter is equally useful in a-m receivers for interference rejection. Why? Because the nonlinearity of the a-m detector becomes serious only on weak signals, where this nonlinearity produces noise. On weak signals the audio signal-to-noise ratio is considerably degraded from that which prevailed in the i-f amplifier before detection. This weak-signal degradation doesn't occur in an a-m receiver under normal interference conditions, since then the signals are strong enough to operate the a-m detector above the degradation point.

However, you can't have everything; and even with an ssb receiver, a very narrowband filter in the speaker leads won't prevent a strong adjacent signal from generating strong agc, thus weakening the signal you're selecting with the cw filter. This can be annoying if the strong

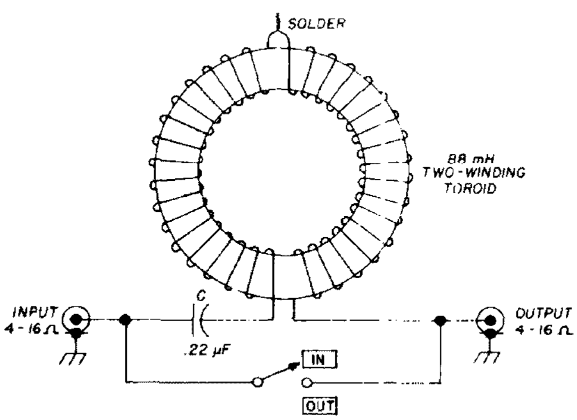



fig. 1. The simplest audio filter. Bandwidth is about 70 Hz at 1.2 kHz; circuit Q is approximately 17.

| C (μF) | Frequency (Hz) |
|---|---|
| 0.22 | 1200 |
| 0.5 | 720 |
| 1.0 | 520 |
| 2.0 | 380 |

adjacent signal is cw, because receiver gain will flip up and down in response to keying. The desired signal will sound like one subjected to severe flutter or fading conditions. The only remedy for this in ssb receivers is a filter ahead of the agc rectifier. In a-m receivers, filtering in the i-f amplifier will prevent this problem.

## circuit Q

The Q and hence the bandwidth of any practical filter is a function of the quality of components and circuit loading. Circuit Q defines the bandwidth as a percentage of the operating frequency.

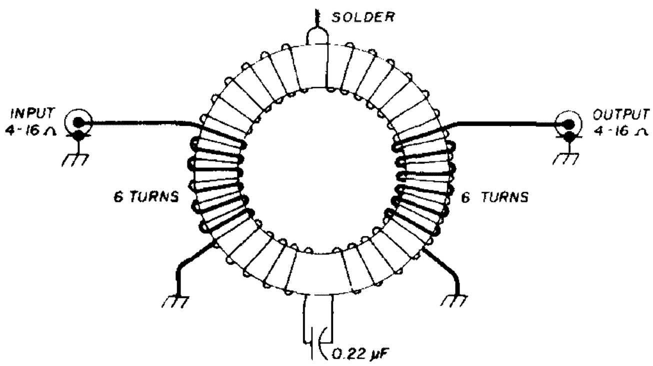



fig. 2. A filter with 40-Hz bandwidth at 1.2 kHz.

Thus a filter with a Q of 10 operating at 1 kHz would have a bandwidth of 100 Hz; and at an operating frequency of 100 Hz, the same coil with a larger tuning capacitor would still have a Q of about 10 but a bandwidth of only 10 Hz. Obviously if your desired signal had interference from a nearby signal, the lower you set the beat note, the easier it would be to separate the two signals with a given filter. With an ssb receiver, about 300 Hz is the lowest you'd care to go, because the audio circuits have very little gain below this frequency. You can therefore choose a frequency between 300 and about 1500 Hz, depending upon personal tastes.

## the simplest filter

The simplest filter circuit having all the desirable features just described is the series-resonant type of **fig. 1**. The capacitor for this circuit should be 0.22 μF for a 1-kHz operating frequency, and 1.0 μF for 500 Hz. This filter has a bandwidth of about 70 Hz at 1.2 kHz and proportionally more or less at other frequencies. The

Q of the filter is thus about 17, which consists almost entirely of the coil Q. Since this is the case the Q, and hence bandwidth, won't vary much if the speaker impedance is between 4 and 16 ohms.

For those who might want a very narrow filter but don't want to operate at a beat frequency near 500 Hz, the circuit of **fig. 2** will give about 40 Hz bandwidth at 1.2 kHz.

The 88-mH toroid in these circuits is an unplotted type available from surplus, and the low-impedance windings of **fig. 2** are each 6 turns of wire wound over the existing windings. Put the input winding on one side of the core and the output winding on the opposite side for the narrowest bandwidth. However, neither the number of turns nor their placing is very critical.

## selectable-frequency filter

As I mentioned at the beginning of this article, these filters are designed for bare-bones simplicity. However, if you'd like more versatility at the expense of a few more components you can make a number of tuning capacitors switch-selectable. A selection of filter frequencies will allow the cw operator to copy signals that may otherwise be lost in extremely heavy interference. A schematic of a multiple-frequency filter is shown in **fig. 3**.

## insertion loss

The insertion loss of all filters of this type is about the same: approximately 20 dB. This means 1 volt out for 10 volts in. This sounds high but really isn't. The audio-volume control can be cranked up to compensate. This insertion loss results because we're using a practical toroid instead of a perfect inductor. The Q of 17 means the coil has an effective ac resistance, including wire loss and core loss (and speaker load) of about 35 ohms. The insertion loss is the ratio of coil ac resistance to load resistance, which can't be improved for that particular coil. The speaker I used was a 3.2-ohm unit, which gave the 10:1 ratio of load resistance to total resistance reflected in the 10:1 input-to-output voltage readings in my measurements. This is explained so you won't feel something is lacking to cause such an apparently large insertion loss.

The filter in **fig. 1** doesn't have dc continuity. Some ssb receivers, with no output transformer in their audio power amplifiers, require a dc path through the speaker. My Galaxy V is an example. A 4- to 16-ohm resistor across the input to the filter will solve this problem.

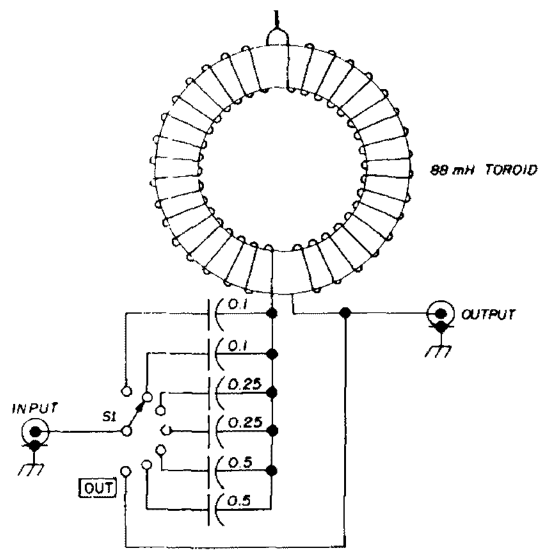



**fig. 3. Frequency-selectable filter for improved cw reception. S1 is a shorting-type switch that connects the capacitors in parallel.**

You've probably seen more complex filters for cw, but none will outperform these circuits *in actual use.* The loss of these filters increases 6 dB for each 40- or 70-Hz excursion from design center frequencies, and response is 40 dB down 200 or 350 Hz away.

## a closing note

The simplicity of the filter will allow it to be placed almost anywhere in your rig. The toroid measures about 1 by ½ inch, and the capacitors can be physically small. The filters will perform as stated if wired into your speaker leads if you'd rather not disturb the wiring of an expensive transceiver.

**ham radio**

# ferrite bead rf stoppers

In certain applications ferrite beads are excellent replacements for rf chokes

Carl C. Drumeller, W5JJ, 5824 N. W. 58 Street, Warr Acres, Oklahoma 73122

Radio-frequency chokes have certain undesirable characteristics that must be recognized and corrected if they are to operate as intended. Rf chokes have "holes" in their frequency-versus-impedance response that can cause resonance in conjunction with stray circuit capacitances.* This phenomenon can result in high circulating currents that will destroy the choke coil.

This article discusses ways of impeding the flow of rf current by using ferrite beads. These beads are made of ferrous particles imbedded in a ceramic material, much like the cores in some rf coils. The physical structure of the beads is somewhat different, however, in that they have a hole through their centers to accept a wire.

## characteristics and uses

Running a wire through a ferrite bead greatly increases the inductive reactance of that length of wire. This reactance follows the familiar 6.28fL law, which shows that as the frequency increases, so does the reactance. At 50 MHz, for instance, one inch of wire through a ferrite bead may show an impedance of 50 + j45 ohms. String on more beads, and the impedance goes up. It increases in a smooth and totally predictable manner—no holes, no peaks.

Now let's see how these beads can be used in ham equipment. For decoupling dc and ac power leads, they're ideal: small, effective, free of dc (or low-frequency ac) resistance, and not susceptible to resonance from associated capacitance.

The increased impedance offered by the beads at higher frequencies suggests

*An example is the requirement for modifying the plate-feed choke in high-power tetrode amplifiers when they first became popular. This is explained in George Grammer's "Pi-Network Tank Circuits for High Power," *QST*, October, 1952 and in the 1953 edition of the *ARRL Handbook*. editor.

another application: that of vhf or uhf parasitic-oscillation suppression. Remember that the impedance consists of a resistive as well as a reactive component. This acts as a potent suppressor of vhf or uhf oscillations, while having very little effect on the primary signal.

This same characteristic can be a temper saver in eliminating rf pickup and rectification of amateur signals by audio equipment. One of the most frustrating jobs you can tackle is that of curing rf pickup (and rectification) in a neighbor's transistorized hi-fi audio equipment. The low impedances encountered in these sets make rf bypassing almost a hopeless task. A ferrite bead or two slipped over the base (or gate) lead of each susceptible transistor can save much hair tearing.

## drawbacks

There *must* be some negative factors! For one thing, the beads saturate with too much current through the wire upon which they're strung. This means you can't use them to decouple the filament of a high-powered transmitting tube in the grounded-grid circuit. Also, they're not suitable in place of an rf choke in parallel-feeding a transistor in a powerful transmitter.

One reason why these handy little devices haven't found greater use among radio amateurs is that they are not listed in most supply catalogs.* Ferrite beads are manufactured by Stackpole Carbon Company, Electronic Components Division, St. Marys, Pennsylvania, and by Ferronics, Inc., 66 North Main St., Fairport, New York 14450. You may have difficulty in purchasing small quantities from these sources, however.

In summary, the use of ferrite beads offers a quick, easy, and painless cure for many of the problems confronting the builder of ham equipment. Use them—you'll like the way they work.

**ham radio**

*One exception to this is World Radio, 3415 West Broadway, Council Bluffs, Iowa 51501. WRL lists them on page 82 of their 1970 catalog as number 75A054; 12 beads in a package for $2.00.

# ideas for an electronics workbench

Like to build your own gear? Try these suggestions for an efficient and safe work area

Jim Ashe, W1EZT

What would you say is the one most important piece of equipment in your electronics workshop—oscilloscope? Vtvm? Clearly it's your workbench. A well-designed workbench can mean the difference between pleasure and drudgery if you spend long hours experimenting and building your own ham gear. If you're planning a new workshop or would like to improve your existing one, the suggestions offered in this article will help you design a work area having an efficient arrangement of facilities. Design suggestions also stress comfort and safety.

## basic requirements

Key requirements for an electronics workbench are accessibility to test equipment, tools, and spare parts; adequate strength; and reasonable cost. If you work on modern equipment, as I do, you probably won't have a piece of equipment heavier than 50-75 pounds, so a massive bench is not a requirement. The trend of modern electronics equipment is such that a good workbench can be built from scratch for less than it can be purchased. Let's take a look at some design considerations.

## size and height

A bench with dimensions less than 1½ x 4 feet is probably too small for maximum comfort and efficiency; if more than 2½ x 5 feet, it's really too large for modern electronics work.

The bench working surface can be either 29 inches from the floor if an adjustable office chair is used (recommended for long-time comfort), or 36 inches high for standing at the bench or sitting on a tall chair. (The Sears catalog has a good choice of such chairs.) Whatever the bench height, a slight adjustment will be necessary if you're taller than about 5½ feet. Alternately sitting and standing, over long periods of time, is restful.

The lower height is nice because everything is closer to the floor. For example, a scope on a cart[1] will free much bench space. If the cart is close to the floor, the

low c.g. is insurance against toppling a piece of expensive test equipment. On the other hand, a higher bench will allow several storage shelves underneath. To resolve this problem, I suggest you make some scaled sketches then mock up some experimental designs with cardboard cartons.

## shelving

Shelves are necessary in any workshop, of course. At least one should be mounted over the working surface. Mount it

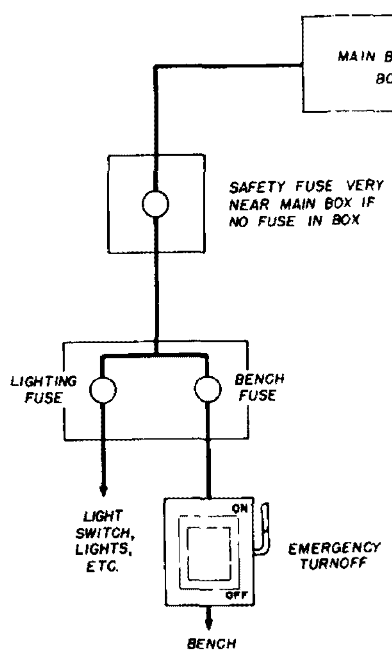



**fig. 1. Simple but effective input wiring circuit. Lighting and power circuits are independent.**

about 15 inches to the right of the working area if you're right-handed; opposite if you're left-handed. It should be about 12 inches above the working surface. If you do a lot of experimenting, you can't have too many shelves. I'd suggest planning your workbench installation so you can add more shelves later. You can buy metal brackets at Sears that will accept a wide variety of plain pine boards.

## parts storage

An excellent source of plastic storage boxes for small parts is Allied Electronics.* These boxes are made by Vlchek. Examples are Vlchek P812 (12 compartments) and P824 (24 compartments) at about $2.00 each. Many arrangements are available for your needs. These little boxes can be nested in a modular arrangement for easy access to parts.

## construction notes

Many facts taken for granted by carpenters and cabinet makers will come as a surprise if you're not familiar with woodworking techniques. For example, if your plan calls for a length of common 2 x 4 lumber, you'll discover that the item you buy is about 1-5/8 x 3-5/8 inches. Common pine or fir is good and inexpensive, but lookout for knots and warped material.

You can save yourself the trouble of cutting material to size for a little extra money. Most lumber supply dealers offer a cutting service. But watch out—if the dealer has one of those large vertical frameworks that takes anything from small board to a sheet of plywood, chances are the salesman will cut the piece by pulling a rotary saw down it. I've seen three of these arrangements, and none would cut square to either edge or surface of the work.

## assembly

How do you assemble the pieces? Again, assuming you're not much of a carpenter, here are some useful hints. Use screws to join large subassemblies so they'll come apart for moving. Use nails for smaller work (shelves, bench top, etc.). An effective and strong joint can be made by running a thread of glue along the material to be joined then nailing the pieces together. The nails should come out without weakening the joint, but it's best to leave them in.

## lighting

Good planning for a workbench includes lighting. Fluorescents are preferred to incandescents, as they offer more than twice the light per hundred watts of

*Allied Electronics catalog 280, p. 267.

power. If a light is poorly positioned, your vision becomes less effective and you tire rapidly.

Three to five well-positioned 40-watt fluorescents will provide good illumination for most work if the fixtures aren't too far from the bench. Used fluorescent fixtures are often available from electricians and should require very little repair. Electrical noise is a problem with some fluorescents; but in industrial and personal experience, I've never found corrective measures necessary.

## power circuits

Your workbench should receive electric power from at least two circuits (see **fig. 1**). A large fuse near the main breaker box is good insurance against unforseen problems. The input line should be run into a box equipped with two more fuses; one for lights and one for power circuits. With this arrangement, the lights won't go out if your work blows a fuse. Note that the emergency switch turns off bench power but not the lights.

## grounding

Grounding *is important.* A good ground system consists of a heavy copper wire (about no. 8) from the workbench to a copper-clad rod driven into the earth. Maximum length of the connecting wire should be not more than about 15 feet.

Avoid transformerless ac-coupled power circuits. They're dangerous and not to be trusted, because a simple accident could put some of your gear at 115 Vac with respect to ground.

## antenna simulation

An antenna connection is a useful but not necessary provision at your workbench. If you can afford it, a signal generator producing a controllable microvolt signal is preferable, because it can be used for realistic sensitivity measurements. I've seen usable gear priced as low as $15.

To approximate operating conditions, a simple dipole can be erected. Make each leg about 0.1 wavelength and connect a 47-ohm resistor across the feed point. The antenna can be strung out of the way in a corner of the shop. Run the coax cable to a connector on the bench.

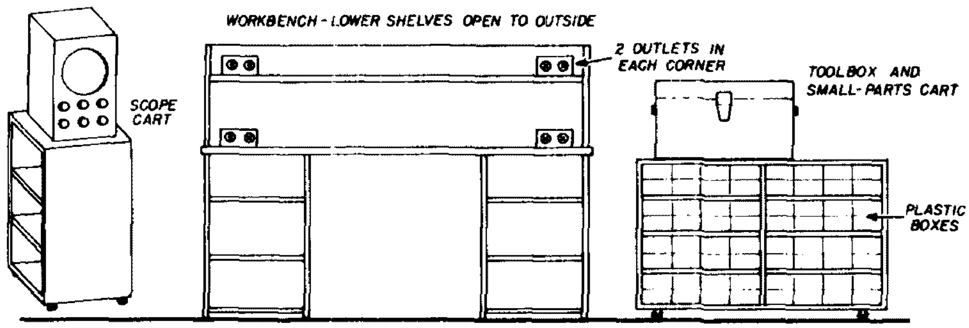



**fig. 2. Suggested workbench installation.**

## summary

A good bench setup is shown in **fig. 2**. A few instruments can occupy the bench working surface, and parts are easily accessible on the shelves below. Test signals are available from a small antenna, and a separate connector can be used to patch a cable to your operating position to pick up your main antenna through an swr meter or tuner. The toolbox, sitting on a small cart, is convenient. Good-quality casters should be used on all the furniture shown in the sketch.

### reference


1. Jim Ashe, W1EZT, "A Cheap & Clean Scope Cart," *CQ,* April, 1970, pp. 34-36.


**ham radio**

# introduction to thyristors

## How to use thyristors, four-layer diodes, silicon-controlled rectifiers, and triacs in electronic equipment

Paul Franson, WA7KRE, 117 Central Street, West Concord, Massachusetts

**Of** the three great basic classes of semiconductors — diodes, transistors and thyristors — there is no question that thyristors are least known to hams. There is good reason for that besides the fact that thyristors have become popular only fairly recently: thyristors have limited application in most ham equipment. Nevertheless, they do have many interesting uses in amateur gear and in the many other facets of electronics most hams enjoy. For this reason, all hams should know something about them. This article is an attempt to present some basic facts about these important devices and how they are used.

Thyristor is a hard word to define to someone who doesn't know what a thyristor is. Rather than belabor the point, I will ignore this problem and simply say that there are many types of thyristors, but the most popular are silicon controlled rectifiers (scr's), triacs and four-layer diodes. Other less important members of the family are five-layer diodes, gated bilateral switches, gate-controlled switches, light-activated scr's (lascr's), gate turn-off switches, and turn-on, turn-off controlled reactifiers.

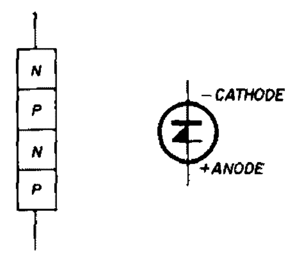



**fig. 1.** Four-layer diode construction and schematic representation.

### four-layer diodes

The simplest type of thyristor is the four-layer diode. Its name comes from its construction. A regular diode is made of two layers of semiconductor material which form the anode and cathode. A transistor contains three: an emitter (cathode), base and collector (anode). A four-layer diode has four layers of alternating p-and n-type silicon; only the outer two layers are connected to terminals.

The symbol for a four-layer diode is derived from the numeral four, as can be seen in **fig. 1**.

The four-layer diode is indeed an odd device. It is a voltage-controlled switch, and has two "states," very high resistance

(off) and very low resistance (on). On the reverse direction, that is, when the cathode is more positive than the anode, it acts like any regular silicon diode; it conducts very little current, and as the voltage is raised, the current increases slowly until a certain voltage (the reverse breakdown voltage) is reached. Then the current increases rapidly. This high current with high voltage drop leads to high power dissipation, which will destroy the diode if the current is not limited to a low value. A four-layer diode is not used in this way; regular diodes are much cheaper!

In the forward direction (positive to anode), however, the four-layer diode acts quite unlike a regular diode. As you

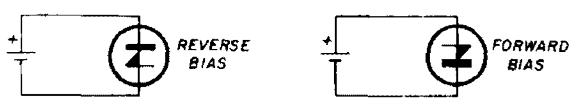



**fig. 2. Forward- and reverse-biased four-layer diode.**

may recall, the forward voltage drop across a regular silicon diode remains approximately 0.7 V over a very wide range of currents through the diode. A four-layer diode, on the other hand, conducts virtually no current until the voltage across it rises to the breakover voltage, when it will suddenly switch to the on state and conduct heavily with a very low voltage drop, much like a conventional silicon diode. Thus, raising the voltage switches the four-layer diode from **off**, (no current flow, or high resistance) to **on** (high current flow, or low resistance). Because of the small voltage drop in the **on** state, very little power is dissipated in the device; a typical drop is 1.2 V at 70 mA, or 84 mW.

If current through the four-layer diode is reduced below the holding current of the diode, as by breaking the circuit, the four-layer diode switches off again. Notice that the four-layer diode is a dc device. This operation of the four-layer diode is shown graphically in **fig. 3**.

While there are many commercial uses for four-layer diodes, ham applications are limited. However, in at least some cases, four-layer diodes can simplify circuitry considerably. One example is the sawtooth or relaxation oscillator; here the four-layer diode acts like a low-voltage, dc neon lamp. Fig. 4 shows the circuit. Since the four-layer diode does not conduct until the breakover voltage is reached, it is out of the circuit for all practical purposes while the capacitor is charging up – until it reaches the breakover, when it suddenly conducts, discharging the capacitor. The cycle then repeats, and the result is a sawtooth output voltage. The resistor and capacitor are chosen to provide the proper repetition frequency. The resistor must be fairly large and the source voltage considerably higher than the breakover voltage for best wave shape. The sawtooth oscillator can be used to generate pulses for triggering scr's or as a code or test oscillator.

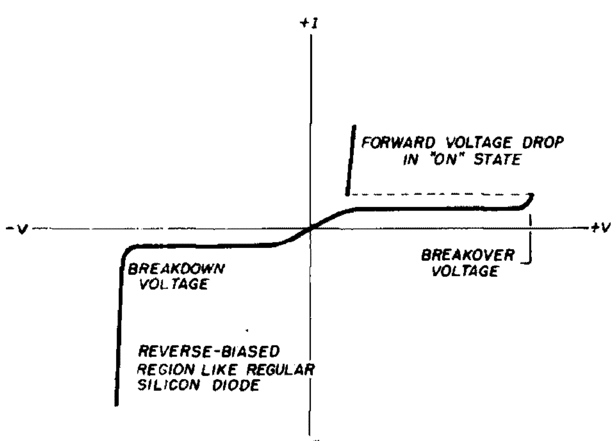



**fig. 3. Volt-ampere characteristics of the four-layer diode.**

Another use of the four-layer diode is as an overvoltage relay, as shown in **fig. 6**. If the source voltage rises to an excessive value, the four-layer diode will trigger, energizing the relay and disconnecting sensitive equipment. The resistor can be used to limit current if the relay coil has insufficient resistance. It will have negligible effect on the switching voltage.

## sources of supply

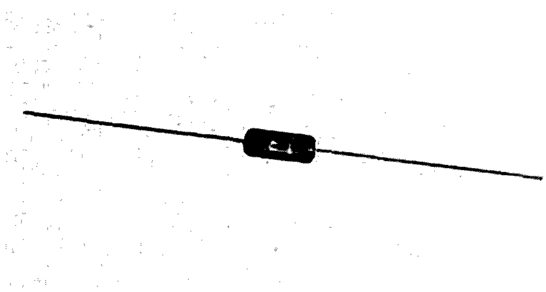
Four-layer diodes are made by a number of companies including Motorola, ITT and Crystalonics. They are fairly expensive to individuals because they are not used very widely and are tricky to make. An example of a commercial four-layer diode is the Motorola 1N5158, which has a nominal break-over voltage of 8 V, a holding current of 1 to 20 mA, and a continuous forward current of 180 mA. It costs about $3.75.

For experimenting, a low-voltage scr can be used as a four-layer diode by ignoring its gate. This is especially nice because low-voltage scr's are undesirable for most amateur uses and can be bought at very low prices. This type of two-terminal operation of scr's is not recommended by manufacturers, though. It can cause a change in characteristics or even damage the device. This will probably not bother most experimenters as long as they stick to low-voltage and low-current operation.

## finding the breakdown voltage

Of course, you'll have to determine the breakover voltage when you use an scr as a four-layer diode. Fig. 7 shows how to find this breakover voltage. Simply increase the variable voltage supply (with an eye on the voltmeter)

Package commonly used for four-layer diodes is the same as that used for conventional diodes.



until the current through the diode suddenly increases and the voltmeter reading drops. The voltage at which this happens is the breakover voltage. The current-limiting resistor should be chosen to keep the current through the four-layer diode and meter to a safe level. For a level of 50 mA, a reasonable value for

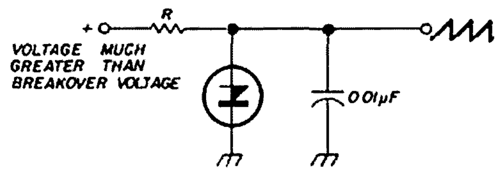



fig. 4. Sawtooth generator circuit.

most four-layer diodes, the resistance should be the maximum voltage divided by 0.05. For 50 volts, it should be 1000 ohms.

This same test circuit can be used for testing four-layer diodes, too. Reject

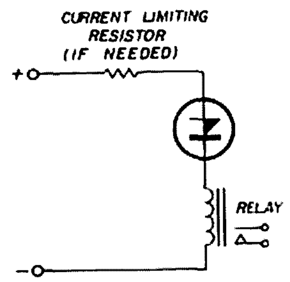



fig. 5. Overvoltage cut-out relay.

four-layer diodes can be bought from some surplus dealers, and it's always a good idea to check them before use.

Another way to turn on a four-layer diode is shown in fig. 6. Here a positive pulse turns the four-layer diode on and energizes the load. This arrangement is

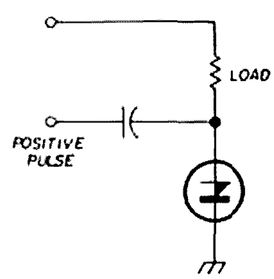



fig. 6. Alternate way to turn on a four-layer diode.

better in many switching applications than using the four-layer diode as a true voltage-controlled switch. The switching point is precisely controlled.

## silicon-controlled rectifiers

The silicon-controlled rectifier, or scr, is the most popular thyristor. Unlike the four-layer diode, the scr has hundreds of practical uses in consumer, industrial, military—and amateur—equipment and is widely used. Modern plastic-encapsulated scr's are inexpensive enough to be used in low-cost appliances, and this, in combination with their usefulness, is revolutionizing the whole idea of control circuits.

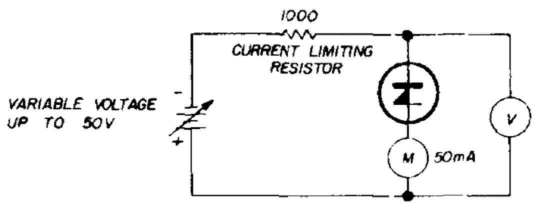


fig. 7. Determining the breakover voltage of a four-layer diode.

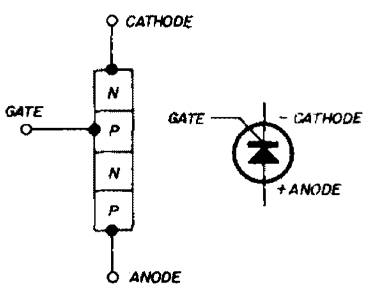


fig. 8. Construction and schematic symbol for the silicon-controlled rectifier.

The scr is very similar to the four-layer diode in construction. It is a sandwich of four layers of silicon containing different amounts and types of impurities. The symbol and construction are shown in fig. 8. The scr can be thought of as a four-layer diode with a connection to the layer next to the cathode. This connection is called the gate, and is what makes the scr so useful.

An scr acts just like a four-layer diode except that it can be turned on when in the blocking state by applying a short pulse of positive current to the gate. The scr will then suddenly switch from the blocking state to the conducting state and have a low voltage drop much like a conventional silicon rectifier. The characteristic curve of an scr is shown in fig. 9. Notice that is identical to the curve of the four-layer diode (fig. 3) except that triggering the gate "overcomes" the forward blocking voltage.

This is oversimplifying slightly, since small values of triggering current actually reduce the forward blocking voltage rather than turning the scr on. However, it's best always to use more than the minimum gate current necessary to switch the scr on and avoid this problem.

The gate trigger current is very small in relation to the amount of current the scr can control. For example, with the 8-ampere 2N4178, a triggering current of 20 mA at 1.5 V will turn the device on. The scr acts like a very sensitive relay with a very high amplification of 400 (8000/20).

In addition to its sensitivity, an scr switches very fast. A conventional inexpensive scr can easily switch on in about 1 microsecond, and off in 20 μs. This means that it can easily switch on and off during one cycle of an alternating current even at 10 kHz (which has a period of 100 μs). This is a very important characteristic of the scrs, and makes it extremely useful for power control. As a comparison, conventional me-

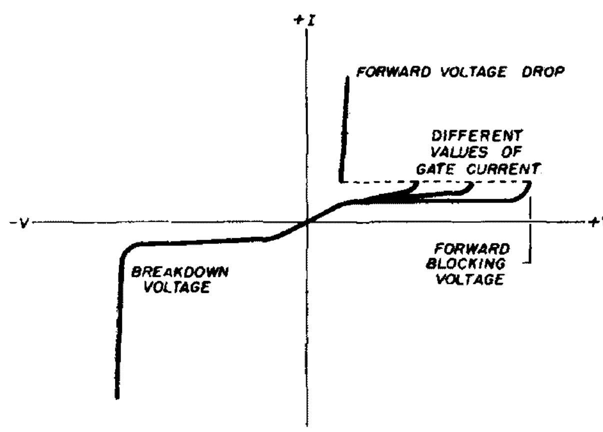
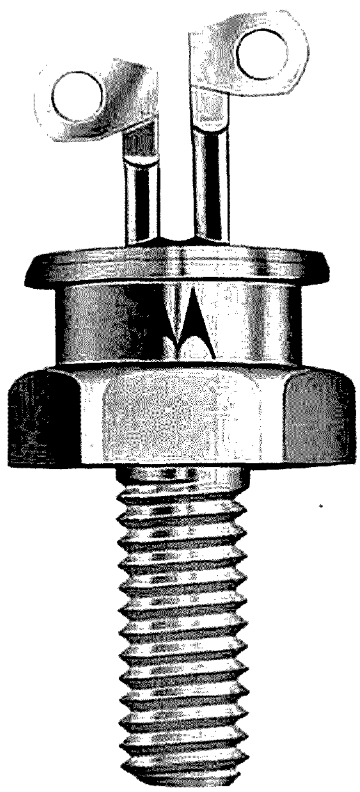


fig. 9. Volt-ampere characteristics of the silicon-controlled rectifier.



Metal-stud package used for the 8-ampere 200V 2N4170 scr and MAC2-4 triac.

chanical relays switch in around 25,000 μs, and very fast reed relays in 1000 μs. Not too surprisingly, scr's are replacing relays in many applications where this speed is important.

Though scr's are often used to replace relays, they have a few characteristics that make them quite different. For one, you can turn an scr off only by reducing the current through it below the holding level. A pulse at the gate turns it on, but you have to break the current (or at least reduce it to a very low level) to turn it off. This is no problem when the scr is used on alternating current since the current falls to zero (and below, to negative values) during each cycle, but some means must be provided to turn off an scr used on dc or it will stay on.

Another important difference is that there is no isolation between the control circuit and the controlled circuit in an scr, as there is in mechanical relays. This may or may not be a problem.

Silicon-controlled rectifiers have many other advantages over relays, however: no contact bounce, no burned contacts, no contact arcing, extremely long life, no movement or acoustic noise and smaller size.

Some typical relay-type scr applications are shown in figs. **10** to **13**. Fig. **10** is a simple latching switch; it operates on dc. Simply applying a positive voltage to the input turns the scr, and hence the load, on. The positive voltage can be taken from the positive supply through a resistor. To turn the load off, power through the scr must be interrupted momentarily. If a conventional relay is used as a load, this makes an excellent latching relay. In any case, a small input signal can control a large load current.

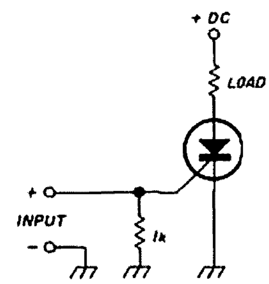



fig. 10. Simple latching switch.

Figs. **11** and **12** are simple lamp drivers (as for use with ic's). In each case, a positive pulse turns on the lamp. In fig. **11**, since the circuit operates from dc, it has a "memory" and will remain on until power is interrupted. In fig. **12**, the lamp does not remain on after the input is removed since the scr is operated on ac. Incidentally, note that in this circuit (but not fig. **11**) the lamp receives power only half the time (when the anode is posi-

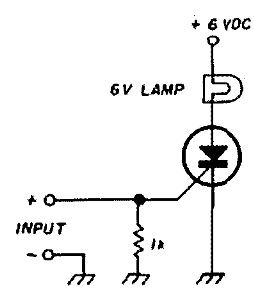



fig. 11. Lamp driver with memory.

tive) so will not light to full brilliance. Because of this, a 4.5-V lamp is suitable for use on 6 V. These last three circuits can be used on high line voltages as well as on low voltage if the proper components are used.

A full-wave ac switch is shown in fig. 13. It applies full supply voltage instead of half to the load. This makes it more practical for most ac applications than the simpler half-wave switches.

These relay-type scr circuits represent a fraction of many others possible. Any ham can likely think of other examples. However, another type of application far overshadows the relay uses: it is in phase control of ac power that scr's really shine.

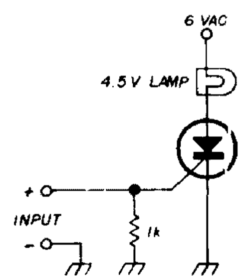



**fig. 12. Lamp driver without memory.**

No other device has proven so versatile in this use. The motor speed controls used in electric drills, blenders and lamp dimmers are common examples known to everyone, but scr's are also vital to modern

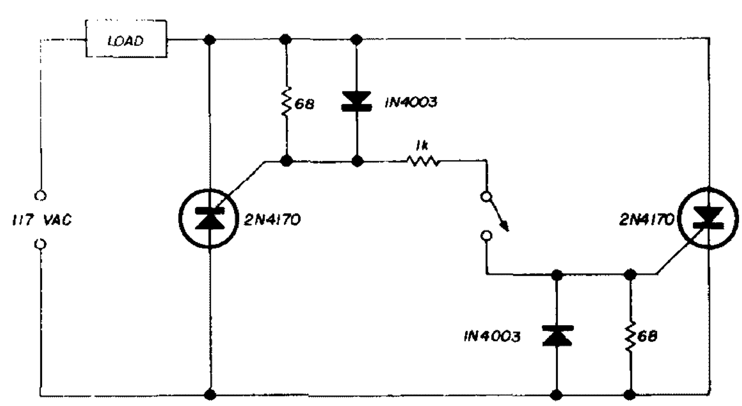



**fig. 13. Scr static ac switch.**

industrial power controls.

Ac phase control is quite easy to follow if you understand peak and rms values of alternating power. Fig. 14 is a

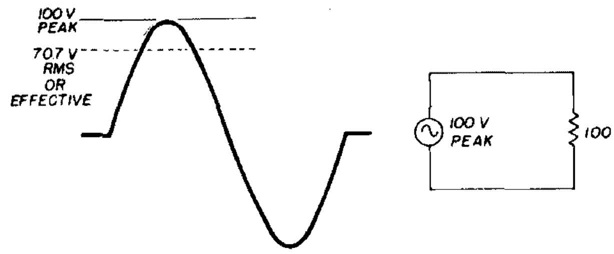



**fig. 14. Peak and effective values of alternating current.**

drawing of a standard sine-wave ac voltage. The peak value is 100 volts and the effective or root-mean-square (rms) value is 70.7 volts. An alternating voltage with a 70.7 V rms value has the same heating value as a 70.7 Vdc. The 117 V line is 117 V rms.

Now suppose this 70.7 V ac voltage were connected to a 100-ohm resistor. By Ohm's Law (I = E/R), 70.7/100 or 0.707 amperes would flow. As a consequence, the power dissipated would be (by P = EI), 70.7 x 0.707 or 50 watts. If we wished to dissipate less power, we could reduce the voltage to a lower value, say to half (50 V peak, 35.35 V rms); the power would be 12½ W (not 25 W, figure it out). However, reducing the voltage is not always that convenient. Adjustable autotransformers such as General Radio Variacs are fairly large, heavy and expensive.

Suppose we took a different approach. Instead of trying to reduce the peak and rms voltages with a Variac, let's leave the peak value alone and reduce the rms, since it's the important one in this application. One way to do this is to keep the

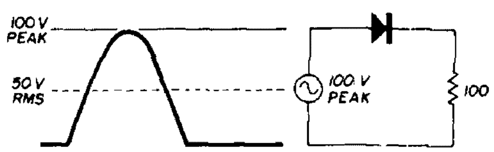



fig. 15. Rms value of rectified alternating current.

voltage from the load part of the time. A diode in series with the load would do this by blocking either the positive or negative half cycles (see fig. 15). Then the

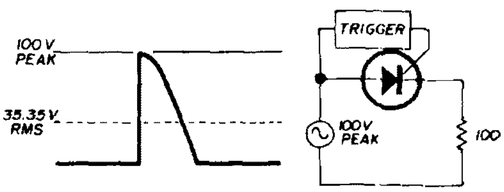



fig. 16. Rms value of alternating current with scr triggered halfway through the positive cycle.

effective value would drop to 70.7 percent of the value without the diode, or to 50 V, reducing the power to 25 W. Why 70.7 percent instead of 50 percent?

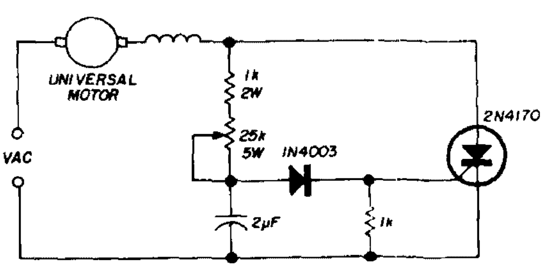



fig. 17. Simple half-wave motor speed control (not for induction motors).

Remember that you reduce the sum (root-mean-sum) to one half, then take a square root.

A diode works well, but gives limited control – full on, 50 percent of power, or off. What is needed is a device that offers better control.

The scr does just this. It can be turned on so fast that a load can be turned on for only a small part of a cycle; fig. 16 gives an example. Here the scr is not triggered on until half way through the positive cycle. Then voltage is applied from then until the voltage drops to zero. The scr acts like a conventional rectifier for the rest of the cycle, blocking the voltage. In this case the rms voltage would drop to half, or 35.35 V, giving 12½ W dissipation.

The triggering of the scr can be adjusted to permit effective power between zero and 70.7 percent. Switching in a diode connected in parallel to the scr, but with reversed polarity, will cover 70.7 percent to full power. Other methods, which will be described shortly, can be used to obtain full variable control without switching.

## practical applications

A practical scr speed control must include some method for triggering the gate on. Probably the simplest practical speed control for universal motors (ac-dc motors) is the circuit shown in fig. 17. It controls the average motor voltage by setting the firing point of the scr. The time required for the capacitor to charge to the gate turn-on voltage is set by the potentiometer. Once the scr is on, the capacitor voltage drops to less than the forward voltage drop of the scr for the rest of the half cycle. During the reverse half cycle, the diode blocks any current which would try to pass through the gate. The capacitor is then ready to begin charging when the next half cycle starts.

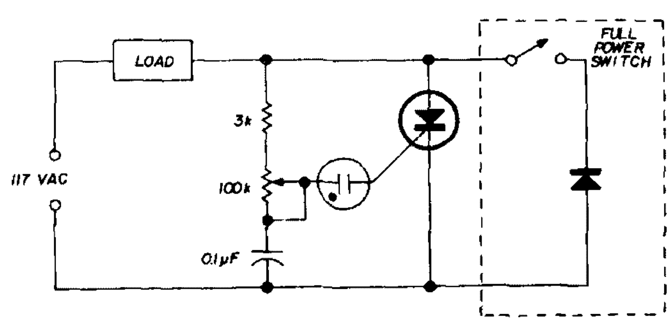



fig. 18. Half-wave phase-control circuit using a neon-lamp trigger.

A better control is shown in **fig. 18.** In this control, the neon bulb acts as a relaxation oscillator and generates pulses that trigger the scr. Adjusting the potentiometer changes the point in the cycle at which the bulb fires, varying the rms voltage applied to the load. Since this is a half-wave control, it can be used only to 70 percent of brightness or speed. Adding the diode and switch fills in from 70 percent to about 95 percent. This circuit is not suitable for control of the ac primary of a transformer because of its imbalance.

Neon bulbs are cheap triggers, but are unreliable and do not permit full range of control. They are also sensitive to radiation: heat, light and radioactivity. This can cause fast changes in speed with changing conditions. The bulbs used in commercial circuits contain a tiny amount of radioactive material that "biases" the bulb and prevents change under normal conditions. If you use a conventional bulb, such as an NE-2, at least paint it black or shield it.

An improvement in the circuit in fig. **17** results from replacing the neon bulb with a "solid-state neon bulb," a bilateral trigger diode (BLT). A bilateral trigger diode has characteristics similar to a neon bulb, but breaks over at about 20 to 30 volts instead of 60 to 70 V. This results in better control. Incidentally, a bilateral trigger is a three-layer diode, and is not a thyristor. It is a symmetrical transistor with no connection to the base. A Motorola MPT20 BLT costs about 65 cents in a single quantity.

A unijunction transistor **(fig. 19)**, four-layer diode, two transistors, or even

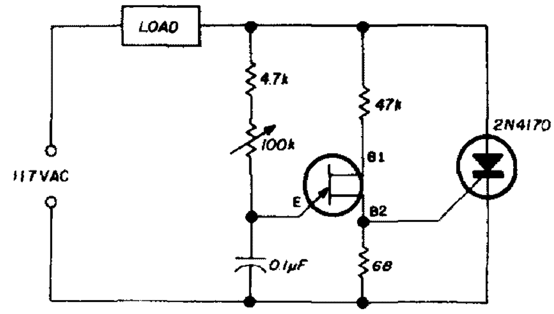



**fig. 19. Half-wave speed control using a unijunction-transistor trigger.**

an ic relaxation oscillator can also be used for triggering an scr.

Throwing the full-power switch is a nuisance, so full-wave control is prefer-

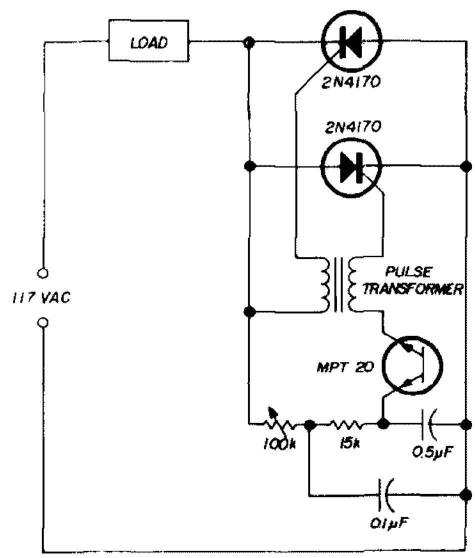



**fig. 20. Full-wave speed control or "variable transformer" using two silicon-controlled rectifiers.**

able in most uses. One way to obtain this is with two scr's connected back-to-back, as shown in **fig. 20.** Triggering is slightly more complex than might be wished for, though, as the scr's must be triggered out

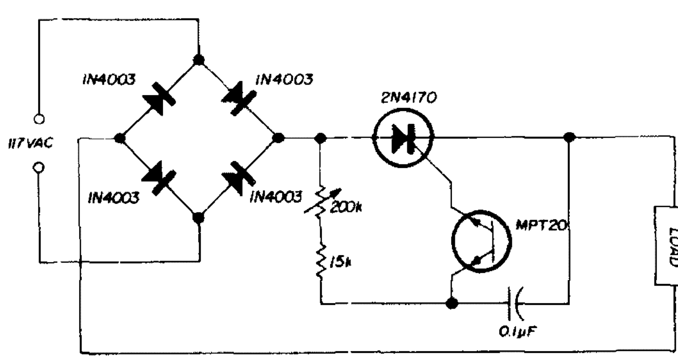



**fig. 21. Full-wave control using one scr and bridge rectifier.**

of phase. One scr is triggered directly (through the primary of a pulse transformer), and the other is triggered by the out-of-phase secondary winding. A simple transformer can be used for this if it can pass a short pulse, i. e., few turns, low capacitance, high-frequency core. Special triggering transformers are made by Sprague and others.

Another approach to full-wave control uses one scr and a simple triggering circuit, but requires a full-wave rectifier bridge (fig. 21). The bridge supplies single-polarity ac to the scr so that it is effective on both cycles. Both figs. 20 and 21 can be used for transformer primary control, i. e., as an adjustable transformer.

A number of other scr projects that have been described in ham magazines are listed in the bibliography.

## what scr to use

Many inexpensive scr's are available to the experimenter. The 2N5060 is a plastic-encapsulated 800-mA, 30-V scr that costs about 69 cents. The 2N4170 is a stud-mount, 8-A, 200-V scr for general line use. It costs about $2.10. A plastic version of the 2N4170 is the 2N4442, which costs about $1.50.

Scr's are also available from surplus dealers. Unfortunately, it requires fairly sophisticated equipment to check them properly, and nothing is more disheartening than having a project not work because of bad or marginal semiconductors.

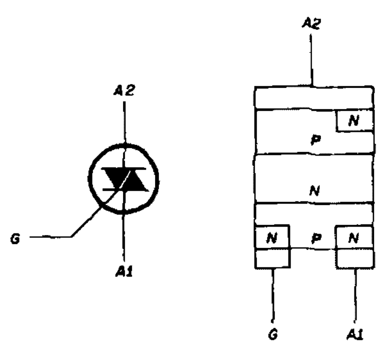


fig. 22. Schematic and construction of the triac.

## the triac

The triac is a three-terminal (tri-) bidirectional thyristor for use on alternating current. Fig. 22 shows the symbol for a triac and its internal construction. The two power terminals of a triac are called anode one (A1) and anode two (A2) rather than anode and cathode since it is not a unidirectional device. The internal operation of a triac is rather hard to understand, but it can be thought of as two back-to-back scr's with a single, more versatile gate. One of the strongest points of the triac is that it can be triggered on in any of four modes: A2+, G-; A2+, G+; A2-, G-; or A2-, G+. The construction of the triac hints at how this is possible. In

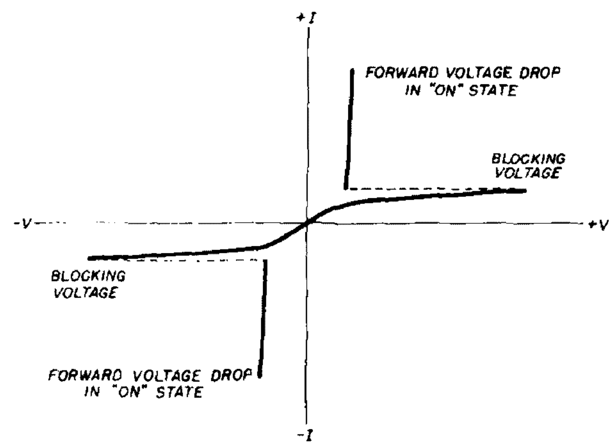


fig. 23. Volt-ampere characteristics of the triac.

each mode, different junctions "disappear" when they are biased properly to form a simpler device.

The triac, unlike the scr or four-layer diode, has no reverse breakdown voltage; instead it has high blocking voltages in each direction (see fig. 23). If either is exceeded, the device will switch into the on state much like an scr. (Incidentally, this makes the triac protect itself from most voltage transients.) In practical use, of course, the device is triggered on by the gate in a manner very similar to an scr.

Unlike scr's, triacs cannot be used at frequencies much above 60 Hz at present.

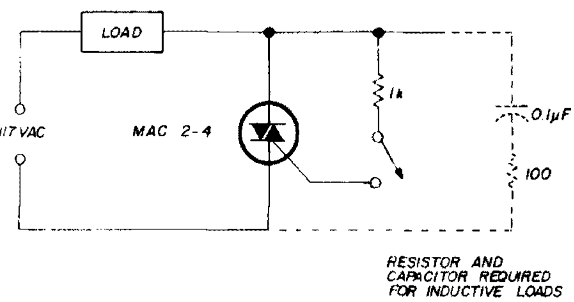



**fig. 24. Static ac switch using a triac.**

Scr's are available with higher voltage and current capability than triacs, and scr's cost less. On the other hand, triacs can simplify circuits considerably and are widely used because of this. A typical triac is the 200-volt, 8-ampere Motorola MAC2-4; it costs $3.45.

## uses of triacs

Triacs have many uses similar to those of scr's. One is as a switch or relay for ac use, as shown in fig. **25**. The resistor-capacitor network is necessary for inductive loads to hold down the rate-of-voltage rise, since an excessive value can trigger a thyristor on.

Fig. **25** shows a simple triac lamp dimmer, or speed control for universal motors. Notice the similarity to the half-wave scr control in fig. **18**, but also remember that this is a full-wave control.

## radio-frequency interference

Radio-frequency interference (rfi), or electromagnetic interference (emi), is often a problem when thyristors are used in phase-controlled circuits. A thyristor can turn on so fast that it creates large amounts of harmonics (hash) up to vhf range. This emi can be conquered, though. Careful shielding and filtering can cure most problems. A completely different approach is called zero-point switching. A zero-point switch turns on a circuit only when the voltage is crossing zero; since no current is broken, there is no emi. To get variable control, a zero-point switching control will apply a few complete cycles of power to the load, then turn off the power for a few cycles. This results in the desired rms voltage level. Zero-point switching cannot be used with lights; it causes flickering.

## conclusion

Thyristors are here to stay in electronics. Though the ham will probably never see too many in his communications equipment, he will run into many in auxilliary gear and household appliances, making an understanding of their working worthwhile.

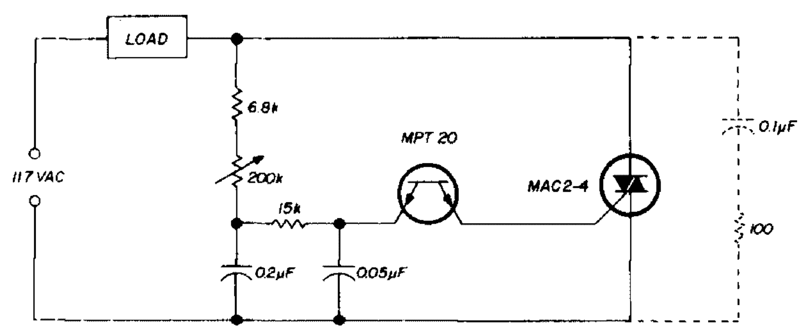
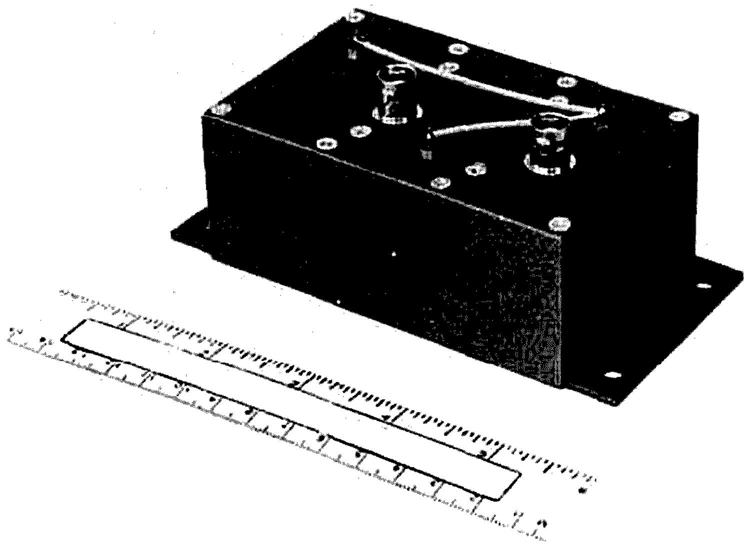



**fig. 25. Full-wave phase-controlled triac speed control, lamp driver, ect.**


### references

1. General Electric, "SCR Manual", fourth edition, General Electric, Schenectady, New York.
2. K. Kokjer, KOJXO, "The Semi-RTTY System," *73*, August, 1967.
3. H. Kukoff, W3HTF, "A Mobile Equipment Protective Alarm," *QST*, March, 1967.
4. R. McCarthy, K6EAW, "A Reactivator for the Semiconductor Activator" (battery charger), *73*, February, 1967.
5. D. Metzger, K8GVK, "The Touch Keyer", *73*, October, 1966.
6. Motorola Semiconductor, "Semiconductor Power Circuits Handbook", Motorola Semiconductor, Phoenix, Arizona.
7. D. Nelson, WB2EGZ, "A Modern Low-Voltage Power Supply with Built-in Short-circuit Protection," *ham radio*, April, 1968.


**ham radio**



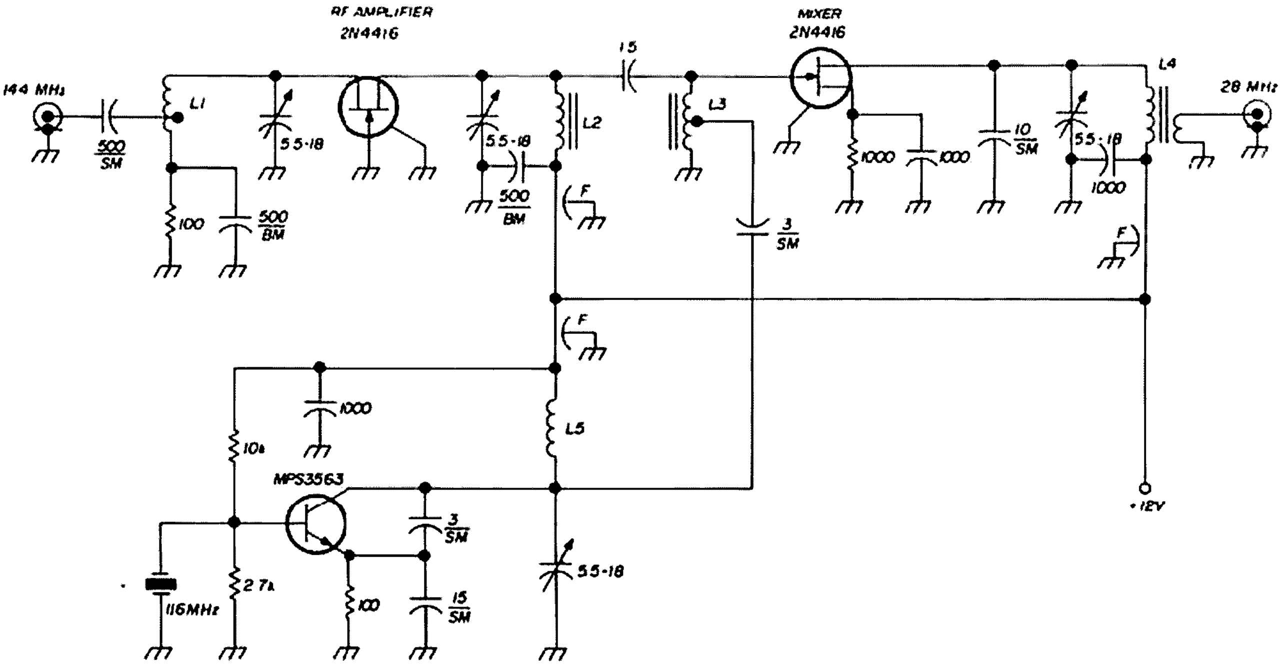
# modular two-meter converter

Like to experiment with new vhf front-end designs? This packaging concept solves many modification problems

Bob Sutherland, W6UOV

New low-noise solid-state devices for two-meter receiver front ends are being announced at such a rate that it's frustrating trying to decide which direction to go. With my limited budget of time, money and ambition it became apparent that I couldn't build a complete converter for every new product announcement. I finally decided to go in the direction described in this article: the modular approach. This allows maximum circuit modification with minimum effort.

The converter described here consists of a bipolar oscillator, jfet mixer, and jfet rf stage (fig. 1). The converter is built on a flexible box-chassis. After the basic circuit is completed, all sorts of low-noise front-end designs can be constructed on



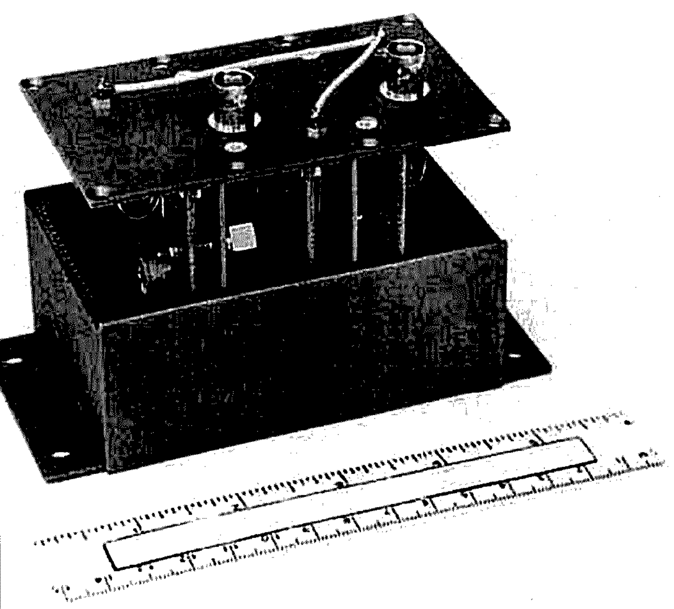

| | | | |
|---|---|---|---|
| L1, L3 | 8 turns of no. 22 enameled wire tapped 3 turns from cold end. Core Micrometals T30-0 | L5 | 3 turns no. 18 wire 3/8" dia. x 3/8" long |
| L2 | 8 turns of no. 22 enameled wire. Core Micrometals T30-0 | F | Gulton MF-223 miniature low-pass filter |
| L4 | Primary 21 turns no. 28 enameled wire. Secondary 3 turns no. 28 enameled wire over cold end of the primary. Core Micrometals T37-10 | | |

All three 1000-pF capacitors are ceramic (Erie type 8133 or Vitramon equivalent)

Micrometals cores are available from Amidon Associates, 12033 Otsego Street, North Hollywood, California 91607

fig. 1. Schematic of the modular two-meter converter. Inexpensive jfet's are used in rf amplifier and mixer; oscillator circuit will deliver 15 mW with a 12-V supply. BM indicates a button mica; SM indicates silver mica.

Completed converter partially inserted into enclosure. Aluminum shields slide into slots on the side of the box, providing good interstage isolation.



individual boards, placed into the proper enclosure, and connected by cable. The final circuit can be updated easily as new designs come along.

## packaging

The circuit isn't new, but the construction technique is somewhat different from that used in the usual two-meter converter. The photos show the modular packaging scheme. The box is a Pomona Electronics model 3306 with a 3328 bottom plate. The aluminum divider for the internal shielding is model 3327. The glass-epoxy circuit boards were etched, as explained later. I think it's evident that any one of these units can be removed and replaced with a new design with little effort. Remaining circuits need not be rebuilt. Also, removed boards are still available for future use and comparison purposes.

## the circuit

The local-oscillator circuit was supplied by WA6NCT as his favorite. It uses a Motorola MPS3563, which is one of the inexpensive epoxy-package transistors. In

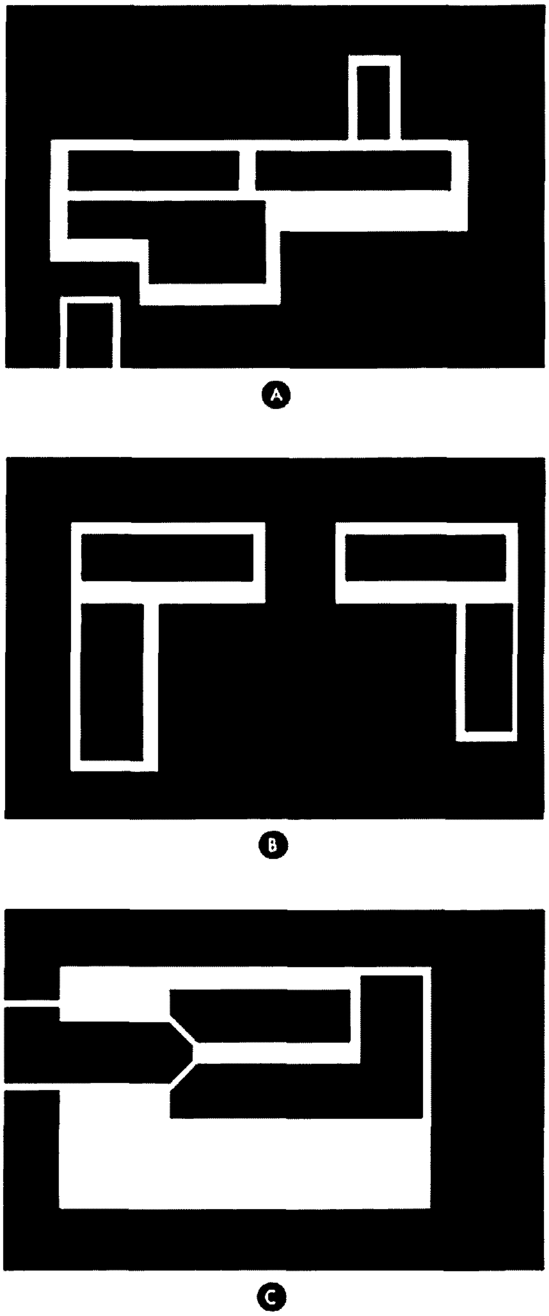
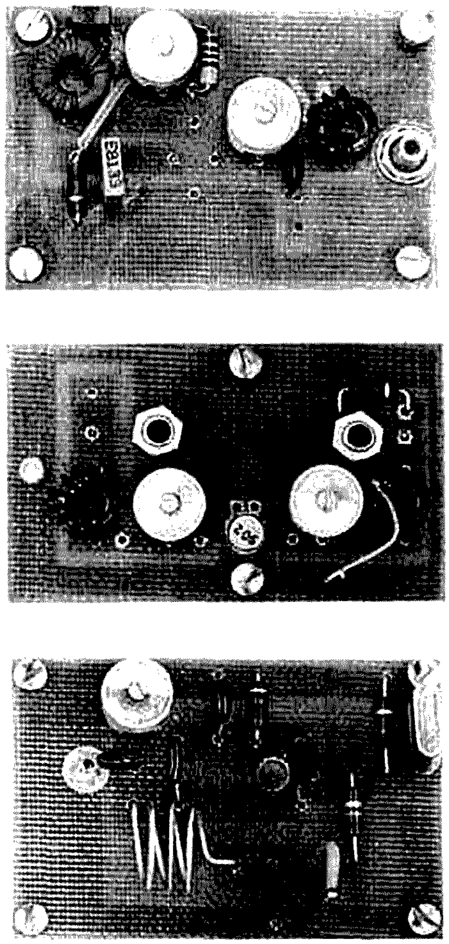


fig. 2. Full-size drawings of circuit boards. Oscillator, rf amp, and mixer are shown in A, B, and C respectively.

quantities from 1 to 99 it sells for fifty-five cents. The oscillator operates directly at 116 MHz and starts every time voltage is applied. Tests were run on the oscillator independently of the remaining converter circuits. With a 12-V power supply, the oscillator delivered 15 mW into a 50-ohm load. The second harmonic was down 27 dB, and the third harmonic was down 44 dB below the

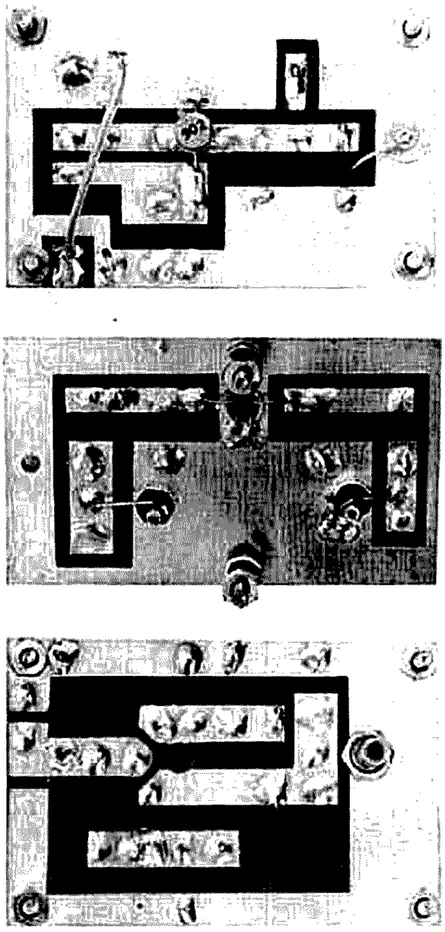


Top view of circuit boards. Top to bottom are the oscillator, rf amplifier, and mixer.

fundamental signal. All other harmonics were better than 60 dB down.

The mixer is a standard 2N4416-jfet type popular with many vhf amateurs. The gate circuit consists of a coil wound on a small toroid and tuned by a ceramic trimmer. The toroid coil has a Q of 135. The toroid was chosen to allow tight packaging of the converter. The rf fields are fairly well confined to the toroid, which helps reduce any tendency toward instability. The output, or i-f, coil is also a toroid. The L/C ratio of the i-f tuned

circuit was chosen to give a 3-dB bandwidth of 2 MHz, and the circuit was peaked for optimum response in the middle of the lower 2 MHz of the 144-MHz band. The converter covers the entire band and has a slight decrease in gain above 147 MHz. The 2N4416 mixer source resistor was chosen to establish the proper bias point.



Bottom view of circuit boards. Oscillator, rf amplifier, and mixer are shown top to bottom.

To determine the correct value of the source resistor, I measured the zero-bias drain current of the 2N4416 actually being used—in my case, it was 11.2 mA. Various values of resistance were tried until the drain current dropped to about 25 per cent of the zero-bias drain current, or 2.8 mA. The local-oscillator injection was then adjusted by the choice of a 3-pf coupling capacitor and the placement of the tap on L3 until the drain current rose to about 40 per cent of the zero-bias drain current, or 4.4 mA. The resulting bias voltage is about 0.7 of the pinch-off voltage, which allows a good compromise between high-conversion transconductance and low cross-modulation with high local-oscillator injection.

The rf amplifier is a 2N4416 in a common-gate circuit. The input and output coils are toroids very similar to the mixer input coil. The bias for this stage is provided by resistor R1. A blocking capacitor from the coax center conductor to the tap on the input coil prevents shorting the bias supply of the 2N4416 rf stage.

## construction

Glass-epoxy single-sided circuit board is used for all three boards. Large areas of copper reduce the lead inductance. Narrow traces are not good practice in the vhf uhf region. The boards are coated with Kodak KPR photosensitive resist. A mechanical negative is made by placing Rubylith over the circuit board pattern and removing the red filter with an Exacto knife from areas where copper is to remain.* An ultraviolet light is used to

* Presensitized board is available from Allied Electronics, 100 North Western Avenue, Chicago, Illinois 60680. Rubylith is available from most graphic arts stores.

"What's this QRN signal you send everytime I come into the room?"

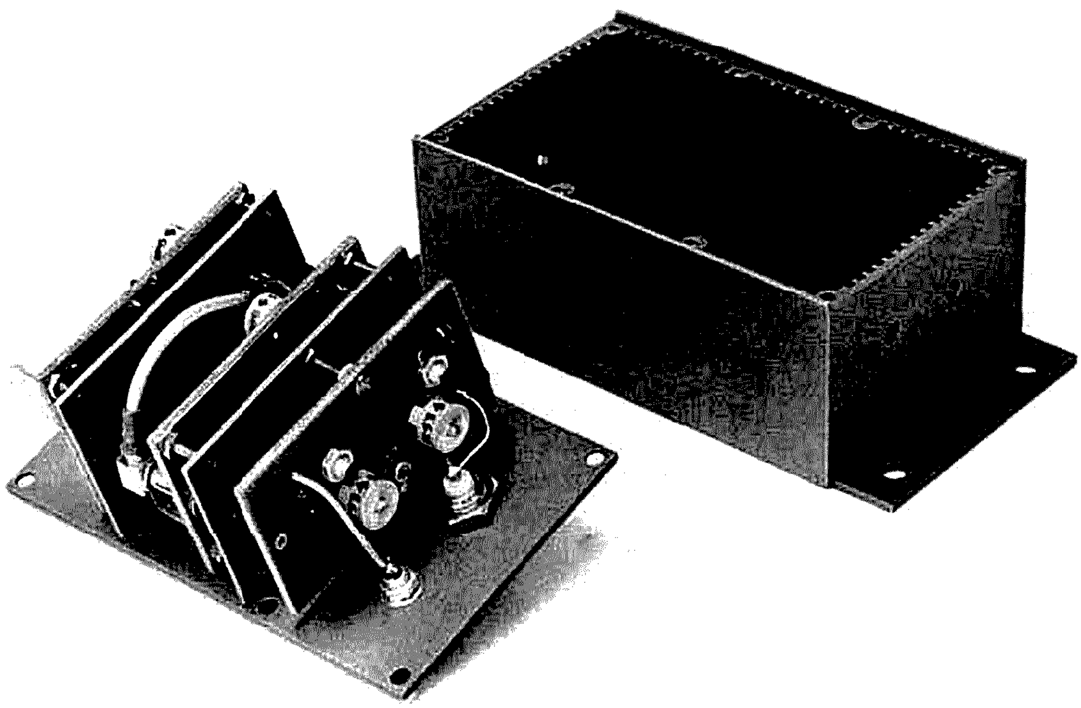
expose the board. Kodak developer can be used, or plain tri-chlorethylene if available. Amateurs with photographic equipment can make a negative from the scale patterns.

Another technique that works well but doesn't yield as good a job is to use ordinary Scotch electrical tape for the resist. Mask off (with the tape) all copper not to be etched. A warm ferric chloride bath can be used for the etchant. The boards to be etched are placed face down in the solution and agitated to speed up the process. After the boards are etched, they can be sheared or cut to size.

Boards are mounted vertically on the box lid. Aluminum shields are placed between each active board. The short length of coax with rather expensive end fittings is not really necessary for good performance. The fittings were available, however, and allow for quick disconnect. A short piece of coax with each end soldered into place would be adequate. Small holes through the box walls allow tuning adjustments to be made.

No sockets were used for the transistors. Also, no shielding was required between input and output of the 2N4416s. The common-gate rf stage provides complete stability.

The converter has sufficient gain to operate directly into a Collins 75A4 receiver, as modified according to a previous article in *ham radio* by W6ZO.[1]



**Interior construction of the modular two-meter converter.**

references


1. Raymond F. Rinaudo, W6ZO, "Improving Overload Response in the Collins 75A4 Receiver," *ham radio,* April, 1970, p. 42.
2. Siang-Ping Kwok, "Field Effect Transistor RF Mixer Design Techniques," Motorola Semiconductor Products, Inc., Technical Information Center, P. O. Box 20912, Phoenix, Arizona 85036.
3. "RF FET Data Packet," Siliconix, Inc.
4. "Field-Effect Transistors in UHF Tuners," Texas Instruments, Inc.
5. "Small-Signal Performance of a UHF Junction-Gate FET," Texas Instruments, Inc.
6. S. Weaver, "UHF Mixer Using 2N3823 FET," Texas Instruments, Inc.


ham radio

# improving the voice commander fm sets

Here are some simple modifications to cure receiver instability and change bandwith response of this equipment in various models

Don Chase, WØDKU, 3315 South Mount Carmel, Witchita, Kansas 67217

**Many requests have come** to my attention for help with the General Electric Voice Commander. Many hams aren't happy with these sets because of receiver instability. This article offers a means for resolving the receiver instability problem and provides information for converting narrowband receivers to wideband operation. A couple of minor but useful transmitter modifications are also described.

The Voice Commander was manufactured in three models: Voice Commander I, II, and III. Model I used subminiature tubes in the transmitter, but the receiver is essentially the same as in the other models. Components and physical layout for the receivers were changed only slightly throughout the entire model series. Models II and III have nearly identical receivers.

An rf amplifier was added to the receiver of the Model III. This circuit is on a separate chassis hidden under the transmitter board. Also included is a ptt relay for an external microphone. In this article, all references are to the Model III; however, they apply to Model II except for the part about the receiver rf amplifier.

The Voice Commanders were manufactured in two production splits. Split 1 covered 132-150 MHz; split 2 covered 150-174 MHz. Most high-split units will tune to 146-147 MHz with no trouble. Information is included here on appro-

priate padding to accomplish this tuning range.

## receiver rf amplifier

The receiver rf amplifier is model 4EA19A10 (low split), or 4EA19A11 (high split). The original unit used G.E. part no. 19C300037-2 transistor. It was replaced in revision A with part no. 19A115413-1 (2N2996). The higher gain of this transistor produced oscillation in some receivers. This led to revision B, in which the collector was moved to a tap on the output coil, and a 10k ¼-watt resistor was added across the coil.

If your receiver is revision A, the 10k resistor is sufficient to tame it. To pad from high to low split, which is normally not necessary, add a 3-pF capacitor across input and output coils.

## curing receiver instability

The basic complaint of receiver instability can be corrected by following the steps outlined below, in which dc voltage distribution is rearranged to eliminate a tendency toward regeneration and oscillation. These changes were incorporated by G.E. into later versions of the Voice Commander. A manual for the Voice Commander III is available from G.E.*

## modifications

The first step is to identify the chassis. Inside the square can are three small chassis with a lid on each. Remove the lids. The chassis are identified as follows, although the numbers are sometimes hard to see.

1. 4EL13A10 (low i-f gain and discriminator board).

2. 4EF29A10 (narrowband) or 4EF29B10 (wideband) high i-f gain second-oscillator/mixer and low i-f filter board.

3. 4EF14A10 (low split) or 4EF14A11 (high split) front-end board.

Begin with the low i-f gain and discriminator board. At each end of the board is a black lead running to the high i-f gain and low i-f filter board. Unsolder the black lead at hole 14 (near the discriminator transformer) and let it hang loose for now. Move the other black lead from hole 13 to the top of R3 (2.2k ¼W) next to hole 13. You're now temporarily finished with the discriminator board.

Remove the high i-f gain and low i-f filter board (center chassis) and turn it over to the solder side. You'll see a wire in sleeving connected from a point in the center of the board (hole 4) to one end. The end connection is at hole 9; remember its location and unsolder the end of the wire from this hole. Move the end of the wire to the opposite end of the board. The black wire hanging there is in hole 14. (You disconnected the other end from the discriminator board.) Remove this black wire and discard. Connect the wire previously removed from hole 9 to hole 14. Install a 2k ¼-watt resistor between hole 4 and hole 9. Use sleeving and dress the leads carefully.

Now turn the board over and locate R2, a 6200-ohm ¼-watt resistor next to hole 13. Install a 0.047 μF capacitor from the top of R2 to hole 13. Unless you have very small components, I recommend that you order the capacitor from a G.E. service center. The part number is 5492638-P6.

Unsolder the black wire from hole 13 and let it hang. This wire connects to the 4EF14 front-end board.

## receiver wideband conversion

If your Voice Commander is a narrowband unit and you wish to convert it for wideband use, you'll need a 47k ¼-watt resistor, a miniature 1k pF capacitor (G.E. part no. 5491500-P7), a miniature 1200 pF capacitor (part no. 5491500-P8), and a miniature 12 pF capacitor (part no. 5495334-P42).

Looking at the top of the high i-f gain and low i-f filter board, locate Q2, which is near the center of the long side and

*General Electric Co., Box 4197, A & SP, Lynchburg, Virginia 24503. Enclose check or money order for $1.00.

next to one of the four slug-tuned coils. (Q2 is a TO-18-size transistor.) Locate the two small holes for the 47k resistor between Q2 and the slug-tuned coil. The 47k ¼-watt resistor will fit into this space, although it's a tight squeeze. In the area bounded by the four coils are seven capacitors. Four of these are 110 pF. Leave these alone. Replace the 3300-pF, 2.7-pF, and 4700-pF capacitors with 1000-pF, 12-pF, and 1200-pF capacitors respectively.

## front-end board

If your unit is one of the high-split sets and you wish to convert it to low-split operation, remove front-end board 4EF14A11. Locate three air-wound coils, and pad each coil with a 4-pF capacitor. Turn the board over, and you'll see a short, black wire hanging from J2. Remove and discard this wire. Install a 5½-inch-long length of black no. 24 wire in J2.

The first-oscillator crystal is interchangeable between Voice Commanders I, II, III, the Progress Line portable, the transistorized portable (TPL), and Voice Director receivers—in case you want to borrow one to check it out.

**"Before you send in that article on converting this rig to solid stage don't you think we should try it?"**

## reassembly and alignment

At this time, replace the three chassis into the square can. Use care, especially with the center board with the added 2k resistor on the bottom. Dress the black wire hanging from J2 on the front-end board through the same slot containing the coax cable. Run the black wire along the edge to the audio-squelch board. (This is a square board with a round hole in the center.) This board will be identified as 4EA18A10 (narrowband) or 4EA18A11 (wideband). Only the purist will worry about the difference here.

Connect the black wire from the front-end board to hole 8, the location on the audio-squelch board already having a black wire going to the low i-f discriminator board. Do not disconnect the wire already attached to hole 8.

The receiver is now ready for alignment. The low i-f is 290 kHz; the high i-f is 8.7 MHz. Test point 1 is the limiter metering, and test point 2 is the discriminator secondary metering. Be careful of the four slug-tuned coils in the low i-f filter, as a strand of rubber is usually placed inside as a friction device and it has a strange feel on the tuning tool.

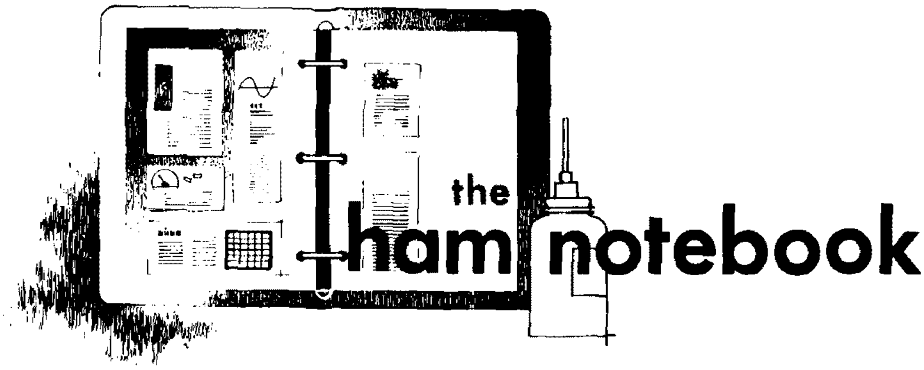
## transmitter

The Voice Commander II and III transmitters are simple and straightforward. On some models, the crystal plugs in; on others it's soldered in. A tuning chart is mounted inside the back cover of all units. The only coils needing alteration are those in the driver and final, and a simple squeeze will suffice.

The audio system of the Voice Commander transmitter is pretty hot. If you wish to speak close to the microphone, reduce the sensitivity by shunting a 390-470-ohm resistor across the microphone cartridge.

With the modifications described in this article installed in your Voice Commander you will have a unit that is a pleasure to use.

**ham radio**




## improved transceiver selectivity

The advantages of operating with transceivers are well established. However, most transceivers on the market lack adequate selectivity for cw reception. I've seen many circuits for narrowing receiver audio bandwidth, but they are either too complicated or entail too many modifications, which reduce the transceiver's trade-in value.

I use a parallel-resonant circuit between the audio-driver and power-amplifier tubes in my transceiver to improve selectivity. Don't be misled by the simplicity of the circuit (fig. 1), as it works very well. The toroid coil, which is available from surplus outlets, and capacitor C form a tuned circuit at the desired audio-tone frequency. A value of C that will give the desired audio frequency may be chosen from **table 1**.

Only two connections need be made at the transceiver—nothing to disconnect or modify. Simply install the tuned circuit between the audio power-amplifier tube grid and ground through a switch. The assembly may be remotely located or mounted at the operating position. The tuned circuit and switch may be mounted in a minibox, with a shielded cable to the transceiver. Thus no holes or other alterations are needed that will degrade the transceiver.

As with all passive networks, this filter causes some loss in audio power, especially as C is increased beyond 0.5 $\mu$F. This means you'll have to crank up the volume control a bit higher than normal. This shouldn't pose a problem, because most transceivers have volume to spare.

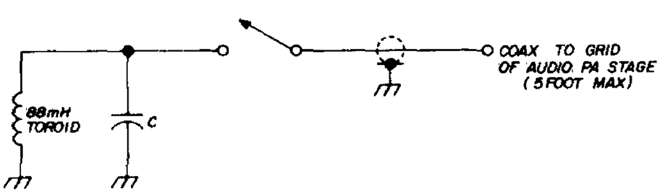


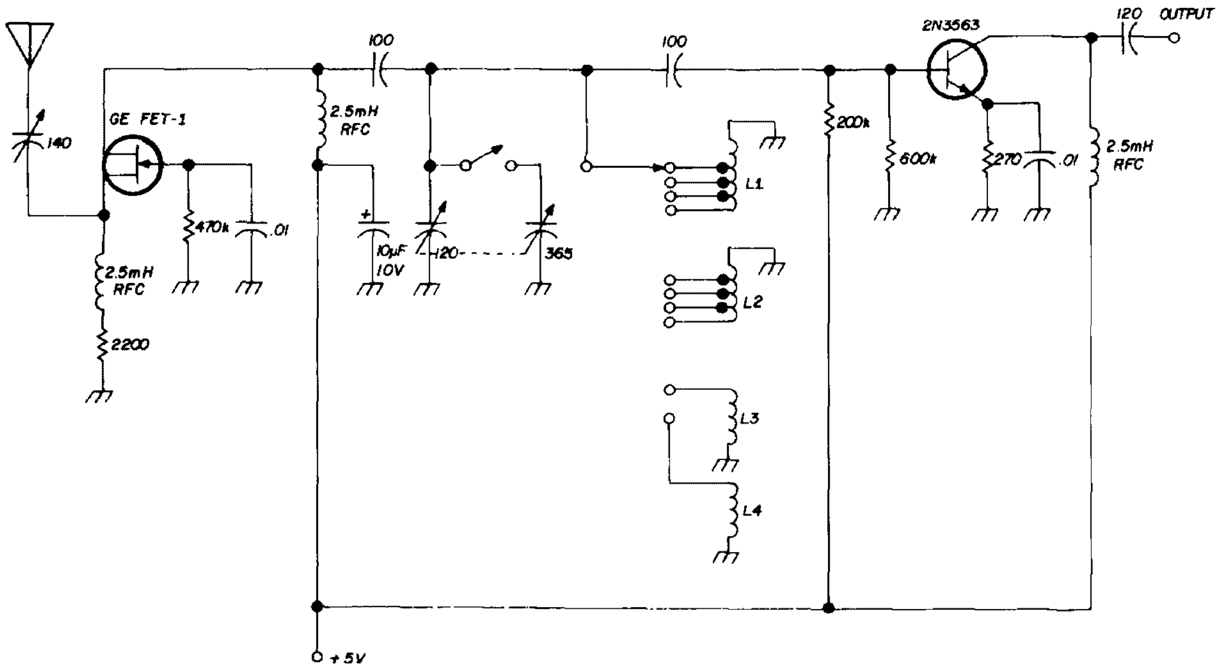

**fig. 1. Simple filter for CW selectivity.**

**table 1. Capacitor values for desired tone**

| tone (Hz) | value of C ($\mu$F) |
|---|---|
| 1700 | 0.1 |
| 1000 | 0.33 |
| 750 | 0.5 |
| 550 | 1.0 |

With this simple filter you'll find that interference from other signals and man-made noise will almost disappear. Cw signals a few hundred Hz apart can be separated easily, and the main tuning dial becomes very sensitive.

**J. Donato, VE3BWD**




| | |
|---|---|
| L1 | 31 turns no. 18 airwound, 5/8" diameter, 1-7/8" long, (B&W 3007) tapped at 2, 11 and 27 turns (tunes 4—40 MHz) |
| L2 | 59 turns no. 18 airwound, 1" diameter, 1-7/8" long, (B&W 3016) tapped at 7, 28 and 38 turns (tunes 1.75—18 MHz) |
| L3,L4 | 80 turns no. 26 closewound on ½" slug-tuned ceramic form, brass slug (L3 tunes 1.05—2.2 MHz, L4 tunes 0.5—1.25 MHz) |

**fig. 2. Preselector for use with general-coverage receivers.**

## general-coverage preselector

The circuit in **fig. 2** was designed for receivers tuning between 0.5–30 MHz. Active devices are an fet in a common-gate, source-input circuit and an npn silicon transistor in a standard common-emitter circuit. The transistors are inexpensive. The fet, a GE FET-1, costs $2.25; the 2N3563 transistor is available from Poly Paks at four for a dollar.

The preselector has fairly uniform gain. Measured at the receiver, preselector gain is 20 dB between 2–30 MHz, with a rising characteristic toward the low end of the broadcast band, where the gain of most receivers seems to be down.

The low-impedance source input of the fet matches low-impedance antennas. The fet is used mainly as an impedance-matching device and has little gain when used alone. High output impedance of the fet and high input impedance of the npn transistor results in low tank-circuit loading; thus tank-circuit Q remains high. With a 5-volt power supply, total current drain is less than 2 mA.

Another fet could have been used instead of the npn transistor; however, the 2N3563'S gain characteristic, together with its low price, made it a desirable choice.

The broadcast band is divided into two segments. A two-gang capacitor, with provisions for paralleling, is used. This gives some flexibility, but it isn't entirely necessary for satisfactory operation. The variable capacitor in the antenna circuit is used to vary input coupling on the lower-frequency bands, since overloading causes cross modulation.

The transistor sockets and related circuit components are mounted on a 2 x 4-inch piece of perf board. This board, plus the larger parts, are mounted on a 4 x 7-inch piece of wood.

With my DX-150, which has spotty sensitivity and some image problems, the preselector improves reception on the low end of the bc band and on 160 meters. With a 25-foot-length of wire for an antenna, the preselector-receiver combination performs very well.

**George Hirshfield, W5OZF**

## printed-circuit labels

Often it's desirable to permanently label components, terminals, and the printed-circuit board itself. These items may be labelled with copper by using dry-transfer letters. The letters resist the metal etchant in the same manner as tape, etch-resist pencils, ink, or paint placed on the copper-clad board.

Dry transfer labels for electronic equipment are made by Datak Corporation, Passaic, N. J. Many amateurs use them to label panels or chassis. The transfers resemble decals, but are more convenient to use. The lables are transferred by lightly rubbing over the characters with a ball-point pen. The sheet of characters is gently lifted to assure a complete transfer. This technique isn't convenient for use with boards using photo-resist chemicals, however.

**Earnest A. Franke, WA4WDK**

## harmonic generator

The circuit shown in fig. 3 will produce 50-$\mu$V harmonics through 1296 MHz with an input of 0.15–1 V from a 100- or 1000-kHz crystal oscillator. With a germanium diode instead of a tunnel diode, harmonics can be heard up to about 147 MHz.

A 2- to 5-ohm bias resistor can be inserted between emitter and ground to prevent possible thermal runaway. As the collector current increases with temperature, the voltage developed across the resistor alters the base current, through dc feedback, to stabilize the collector current. I am not using an emitter resistor, however, and have had no trouble. A larger resistor, up to about 470 ohms, can be used. This will affect the output of the circuit. The emitter resistor should be bypassed to prevent gain loss due to degenerative feedback.

A complete circuit is shown in fig. 4 for those who don't have a 100- or 1000-kHz oscillator.

**Chuck Spurgeon, W5GDQ**

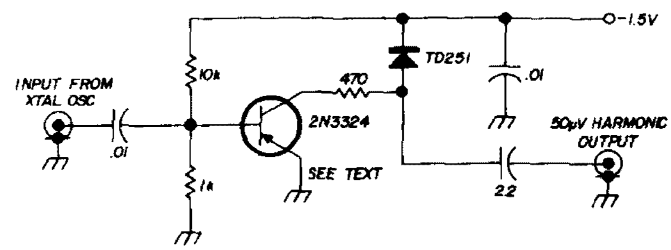



**fig. 3. Basic harmonic generator using a tunnel diode.**

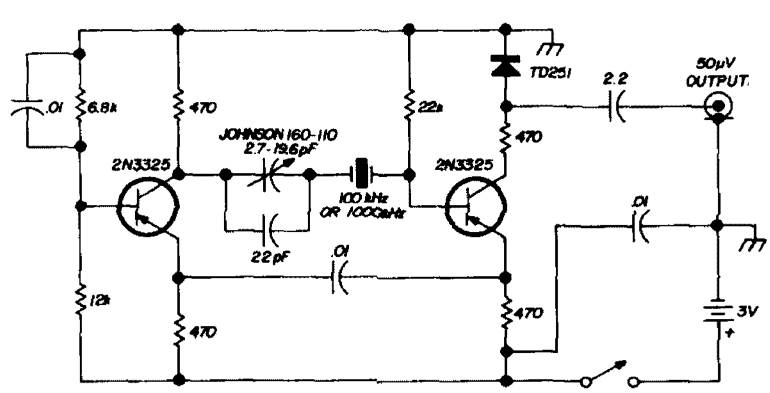



**fig. 4. Complete harmonic generator circuit including reference oscillator.**

## hardware for uhf use

When building vhf and uhf components such as resonant cavities and stripline amplifiers, the need occasionally arises for nylon screws and nuts. Many of us don't have a supply of such items and look for substitutes.

A source of raw material is the flexible plastic handles used on ordinary cotton swabs known (appropriately) as *Q TIPS.* The handles measure 0.10 inch in diam-

eter—just the right size for a 4-40 die. The *Q-TIP* handles will thread nicely for use as screws. All that remains to be done is to saw a slot in one end to accept a screwdriver. The cotton swabs are available for a few cents in any drug store.

Insulating nuts or washers can be made from sheet plastic. Drill and tap a 4-40 hole in the sheet, then punch out the material around the hole with a paper punch. The disc out of the punch becomes the nut or washer.

Ted Swift, W6CMQ

## six-meter mobile antenna

It's easy to assemble an effective six-meter mobile antenna from hardware-store material. An additional advantage is that no holes need be made in the car if the antenna is only temporary.

My antenna was made of two telescoping pieces of do-it-yourself aluminum tubing and an inexpensive plastic ice chest. The plastic foam of the ice chest is the nearest thing to air, and it must be reinforced at critical points. This is important for mobile operation at high speeds.

I mounted the ice chest on a cartop carrier and ran a horizontal dipole through it, fore and aft. Each of two larger sections of the tubing entered the plastic ice chest and were joined in the center by a dowel. The coax feed was run from this point through a hole in the bottom of the ice chest. I inserted the small-diameter tubing in the ends of the larger sections and adjusted the antenna for minimum standing-wave ratio.

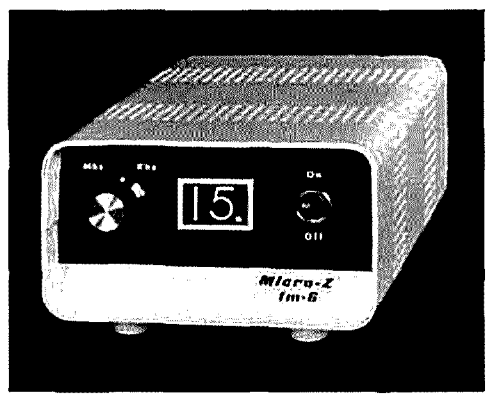
The antenna was highly effective on a cross-country trip. I obtained good signal reports, worked some good DX, and the antenna was durable at high speeds; however, the reinforcement is essential for this.

If a carrier is made from the suction cups and straps available at auto-supply stores, the entire antenna is less expensive than any commercial version and is apparently a better radiator as well. I noticed little evidence of directivity.

Guy Black, W4PSJ

# new products

## digital frequency meter




The FM-6 digital frequency meter, manufactured in kit form by the Micro-Z Company, measures and displays the frequency of any transmitter carrier operating up to 35 MHz—automatically and continuously. Connection is made with a coax T connector and special cable to any transmitter, transceiver, or exciter with an output from 1 to 600 watts. For higher power transmitters, the unit is inserted between the exciter and the final amplifier.

The heart of the FM-6 is a 100-kHz crystal oscillator, adjustable to WWV, that produces a precise gate to permit the unknown signal to pass to a digital

counter and display. To increase accuracy and prevent ambiguous readings, a special dual synchronizer circuit is used to synchronize the gate with the signal to be measured.

The readout consists of two long-life Nixie[R] tubes that display both kilohertz and megahertz together with the appropriate decimal point. As an example, if the frequency is between 7248 and 7249 kHz, the dial will read 7.2 on the MHz scale, and 49. on the kHz scale. On higher frequencies, the highest digit is deleted. (ie. 14.3 MHz would read 4.3). The readout is constant as long as a carrier is on the air. Once every second a sample of the transmitter signal is made, and the frequency reading is "updated." This process takes only a few microseconds and the reading is stable and constant between measurements. Measurement accuracy is within 100 to 200 Hz, and the readout accuracy is 1 kHz.

The unit is all solid-state and uses TTL-MSI (medium scale integration) logic. A high-voltage power supply for the tubes, and a regulated supply for the logic circuits is self-contained in the small metal enclosure (3 1/8" high, 5 1/2" wide and 7 1/4" deep). The easy-to-assemble kit is supplied with all parts, cabinet, circuit board, coax connectors and detailed assembly instructions; $139.50. A factory assembled unit is $169.50. For more information, write to Micro-Z Electronic Systems, Box 2426, Rolling Hills, California, 90274, or use **Check-off** on page 94.

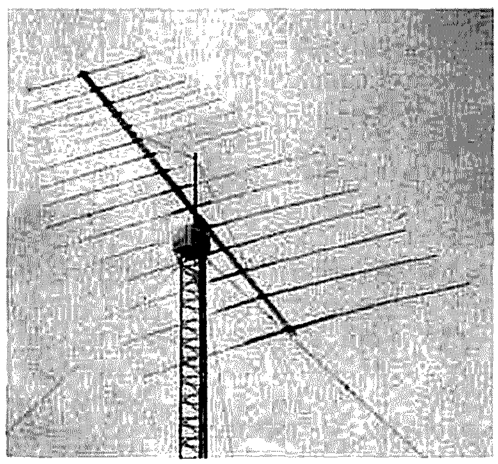
## rotary qsl file

In the new products release in the July, 1970 issue, it was indicated that each rotary QSL holder comes complete with *600* clear plastic pockets. This was in error—each holder comes with *160* plastic pockets. The model CB-8-H rotary QSL file is $8 post-paid from M-B Products & Sales, 1917 Lowell Avenue, Chicago, Illinois 60639. For more information use **Check-off** on page 94.

## low-cost rf detectors

Radiation Devices Company has announced a new line of low-cost coaxial rf detectors for rf demodulation and voltage measurement over the frequency range from 1 to 1000 MHz. Units are available without terminating resistors, or with disc terminations of 50 or 75 ohms. Point-contact or hot-carrier diodes may be specified, with positive or negative output polarity. Frequency response is ±0.5 dB to 500 MHz, and ±1 dB to 1000 MHz. Maximum output voltage is 3 volts rms (point-contact diode) or 25 volts rms (hot-carrier diode). Power dissipation is 1 watt at 25°C. Mounted in BNC-type plug or jack with panel mounting option. Prices range from $12 to $15; for more information write to Radiation Devices Company, Post Office Box 8450, Baltimore, Maryland 21234.

## log-periodic antenna



Hy-Gain Electronics Corporation has announced the availability of an all new rotatable log-periodic antenna that provides continuous coverage from 6.2 through 30 MHz. This antenna has been specifically developed for limited-space applications and is one-half the size of comparable antennas covering the same frequency range. The wideband coverage of this antenna makes it ideal for MARS, maritime, and government high-frequency communications applications.

The vswr over the operating range is

less than 2:1. Gain is 10 to 12 dB over an isotropic; front-to-back ratio is 10 dB average. Power handling capability is 1 kW average, or 2 kW peak. Input impedance is 50 ohms. The boom length is 36 feet, and the longest element is 40 feet. Weight is 250 pounds.

This new log-periodic array, the LP-1017, is priced at $1400. A complete system, the 5017, includes the antenna, Hy-Gain R-3501 rotator and remote-control indicator, and a roof-mounted 20-foot support structure; the 5017 is priced at $2850. For more information, contact the Hy-Gain Electronics Corporation, RR3, Lincoln, Nebraska 68505 or use Check-off on page 94.

## digitone

Digitone Telecommunications Associates offers a number of interesting products for the vhf-fm amateur operator. Digitone decoders, logic processors, and other associated devices have been developed to fill the need for highly reliable, inexpensive solid-state tone-control equipment. The decoding devices are highly versatile in that different combinations may be selected to achieve varying degrees of complexity. Control outputs may be derived directly from the decoder (recognizing single tones), from a binary-decimal converter (recognizing tone pairs), or from the decimal code processor which recognizes and responds to sequential three-digit tone-pair codes.

Products currently available from Digitone include the TDM-202 tone decoder module, the BDC-203 binary-decimal converter, decimal code processors, DPS-201 decoder power supply, COR-221 carrier-operated relay, ACU-222 autopatch control unit, STE-101 single-tone encoder, and TTE-102 teltouch encoder.

For more information on the complete line of Digitone products, and a copy of their latest application notes, write to Digitone, Post Office Box 116, Portsmouth, Ohio 45662, or use Check-off on page 94.

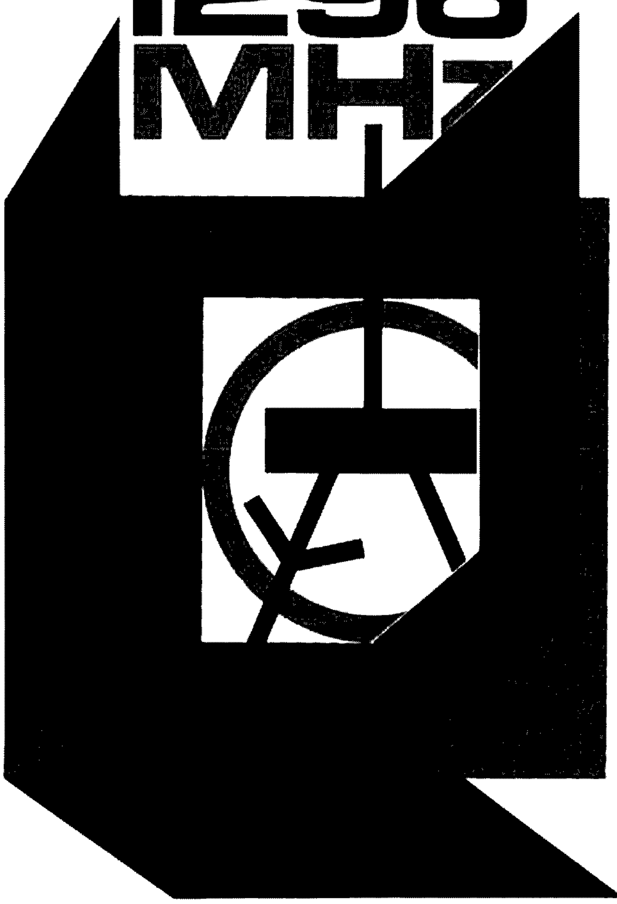
75 cents

focus
on
communications
technology . . .

# ham radio
## magazine

NOVEMBER, 1970

# a solid-state converter for 1296 MHz



## this month

november, 1970
volume 3, number 11



**offices**

Greenville, New Hampshire 03048
Telephone: 603-878-1441

**ham radio** magazine is published monthly by Communications Technology Inc. Greenville, New Hampshire 03048

Subscription rates, world wide: one year, $6.00, three years, $12.00. Second class postage paid at Greenville, N. H. 03048 and at additional mailing offices

Foreign subscription agents: United Kingdom: Radio Society of Great Britain, 35 Doughty Street, London WC1, England.

All European countries: Eskil Persson, SM5CJP, Frotunagrand 1, 19400 Upplands Vasby, Sweden.

African continent: Holland Radio, 143 Greenway, Greenside, Johannesburg, Republic of South Africa

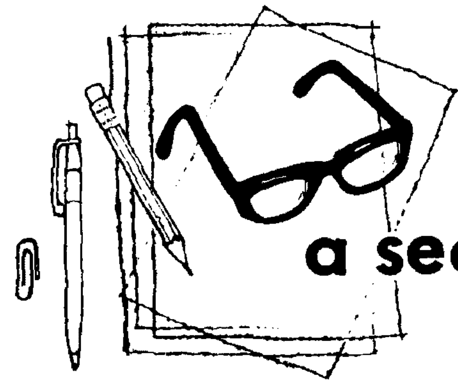

Printed by Wellesley Press, Inc. Wellesley, Massachusetts 02181, U.S.A.

**ham radio** is available to the blind and physically handicapped on magnetic tape from Science for the Blind, 221 Rock Hill Road, Bala Cynwyd, Pennsylvania 19440. Microfilm copies of current and back issues are available from University Microfilms, Ann Arbor, Michigan 48103.

Postmaster: Please send form 3579 to **ham radio** magazine, Greenville, New Hampshire 03048

**ham radio** magazine

# contents



# a second look

by jim fisk

There's a new integrated circuit on the market that could revolutionize future communications-equipment design. The new IC, the Signetics N565 monolithic phase-locked loop, can be used for a large number of interesting applications, including frequency multiplication and division, fm demodulation and a-m detection. Actually, the phase-locked loop concept dates back nearly 40 years, but amateur applications of the scheme have been few and far between – primarily because of circuit complexity and alignment problems.

The basic phase-locked loop consists of a phase detector, a low-pass filter, an amplifier and a voltage controlled oscillator (vco) in the feedback loop. The input signal is fed into the phase detector where it is compared to the output of the vco. If there is a phase difference between the input frequency and the vco, the phase detector produces a dc output signal which is fed through the low-pass filter and controls the frequency of the vco, locking the phase of the vco signal to that of the input. If a crystal-controlled signal is used at the input, the output of the vco has essentially the same stability as the crystal.

The applications for this new integrated circuit are so diverse that the data sheet is already six pages long. For example, you can generate a wide range of frequencies which are multiples or submultiples of the input reference signal; and the output will have the same percentage of accuracy and stability as the input. Tunable vhf converters take on a new light with a phase-locked vfo, as do high-frequency ssb receivers and transmitters.

As a frequency-selective fm demodulator, you can use the phase-locked IC in i-f strips and fm detectors and as high-linearity detectors for very wideband fm applications. As a signal conditioner, you can use this new integrated circuit to synchronize signals, or to track noisy or unstable signals – think of the many uses for this handy device in weak-signal vhf work. The phase-locked IC can also be used for frequency selective a-m detection, or as a coherent a-m detector.

The Signetics N565 also has several functions that are directly applicable to vhf fm operation. Since the circuit can detect tones, you can use it for simple, but effective selective-call systems. If your fm repeater uses multiplex to read out various repeater operational parameters via multiplexed fm telemetry, you can use the integrated phase-locked loop to filter and demodulate the signals.

The frequency range of the N565 is 0.1 Hz to 500 kHz; the more expensive N562 works up to 50 MHz. The circuits will operate with signals of 100$\mu$V to 1 V with best operation at about 5 millivolts. We expect to have a complete applications oriented article on this new device in a coming issue. In the meantime, if you come up with any interesting phase-locked loop circuits, I would like to hear about them.

**Jim Fisk, W1DTY**
**editor**

# solid-state 1296-MHz converter

Introducing a uhf converter from Australia — a trough-line mixer, low-noise pre-amp, and complete portability are featured

H. N. Sandford, VK4ZT, 18 Loch Street, Toowoomba, Queensland, Australia

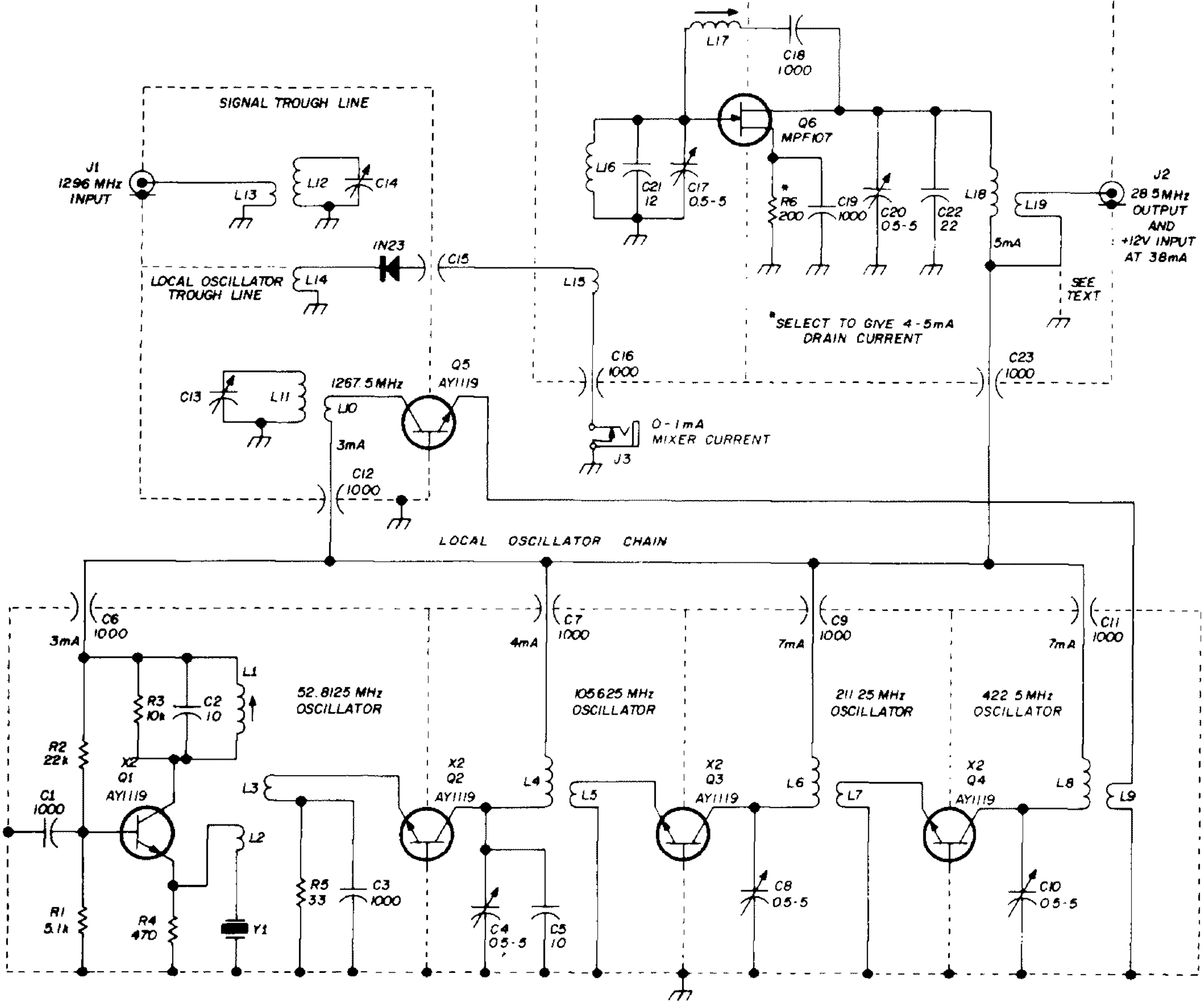
The converter described in this article was used to establish the 138-mile Australian record with VK4KE on 1296 MHz. From the beginning of the project it was decided to develop a solid-state local-oscillator chain to gain experience with transistors in the uhf range. The improved frequency stability would allow narrow-band operation with a consequent reduction in transmitter output power. To obtain any distance on low power, portable operation from 12 V would be necessary.

In the past, the tendency to build units like a battleship stemmed mainly from having access to a well-equipped workshop. As I now use what is probably a typical ham workshop, consisting of a vise and a few hand tools, I was forced to modify construction methods accordingly. The majority of projects are now built using 26-gauge tinplate. It's easy to work, solders well with a 25-watt iron, and most important of all it provides excellent rf shielding. If care is taken with the mechanical design, this light-gauge metal provides adequate mechanical stability. This metal is available at reasonable prices in most cities and is stocked at technical colleges.

## description

The schematic is shown in **fig. 1.** The mixer uses an shf diode in a trough line that feeds an fet low-noise preamp using an MPF-107. The LO chain used five Fairchild AY1119 transistors.* My choice of these devices from an overwhelming number of available types was mainly because of their low cost. I use this

*Type 2N918 and 2N3487 also work well. editor.




C4, C8, C10, C17, C20 0.5-5 pF tubular ceramic trimmers

C13, C14 no. 8 screws with lock washer and nut

C15 diode bypass capacitor (fig. 2 and text)

D1 1N21, 1N23, or other suitable vhf/shf mixer diode

J1, J2 BNC sockets

J3 miniature close-circuit phone jack

L1 12 turns no. 28 on 5/16" diameter slug-tuned form

L2 1 turn single-strand hookup wire, cold end L1 (see text)

L3 2 turns single-strand hookup wire, center L1

L4 5 turns no. 16, 7/16" ID winding length 1/2"

L5, L7, L9 1 turn single-strand hookup wire 7/16" ID

L6 3 turns no. 16, 7/16" ID winding length 3/8"

L8 1¼ turn no. 16, 7/16" ID, winding length 3/8"

L10 1" collector lead of Q5 (Figure 2 and text)

L11, L12 ¼" OD copper tube, 4¼" long, Figures 2 & 4

L13 ½" length no. 18 (see text)

L14 7/8" length no. 18 (see text)

L15 7½ turn no. 28 enam. closewound over cold end L16 (see text)

L16 17 turns no. 28 enam. closewound, 5/16" OD; form mounted over C17

L17 60 turns no. 35 enam. progressive winding on 3/16" OD slug-tuned form (see text)

L18 17 turns no. 28 enam. closewound 5/16" OD form mounted over C20

L19 2 turns single-strand hookup wire over cold end of L18

R6 200 ohm, ¼ watt (select value to give 4-5 mA Q6 drain current)

Y1 52.8125 MHz 3rd overtone crystal

**fig. 1. Schematic of the 1296-MHz converter. A 9.8-dB noise figure is claimed after optimizing the circuits.**

inexpensive npn transistor for most i-f and rf applications where noise figure isn't important. The AY1119 has an $f_T$ of around 450 MHz and will produce 20-30 mW in the low vhf range. The AY1114 is its direct pnp counterpart and may be used when a positive ground is desired.

An i-f of 28.5 MHz was chosen, which allows coverage to 0.5 MHz below 1296 MHz with a receiver that tunes the standard 28-30 MHz band. The third-overtone crystal oscillator, Q1, operates at 52.8215 MHz.

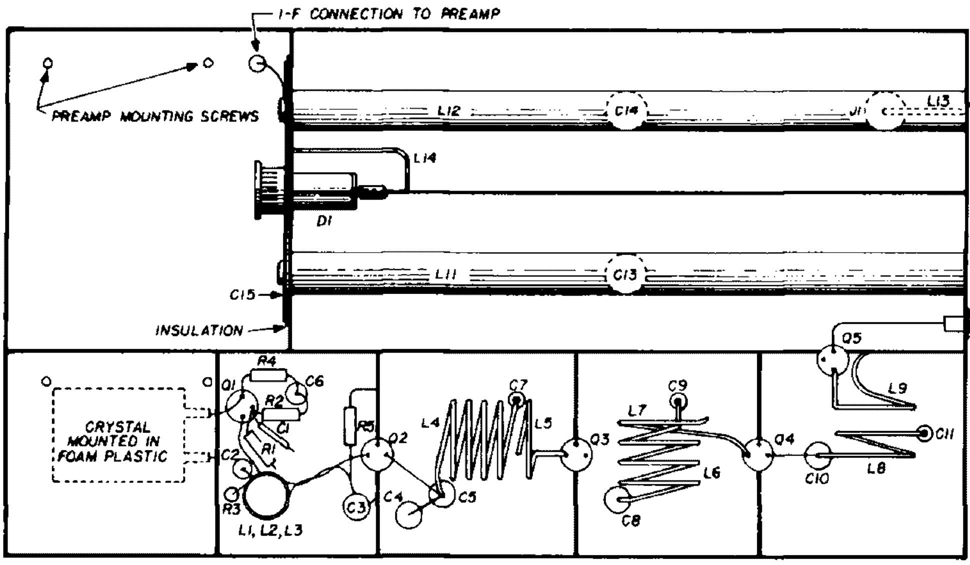
Transistors Q2, Q3, and Q4 are doublers operating at 105.625, 211.25, and 422.5 MHz respectively. Common-base configuration was chosen, as it provides a convenient layout with a minimum of components per stage.

## tripler stage

Transistor Q5 triples to 1267.5 MHz, providing up to 1 mA of mixer diode current. I originally intended to use the more common but less efficient diode multiplier. Fortunately, I didn't have a suitable diode such as the 1N82 on hand. Plenty of output was available from Q4 on 422.5 MHz, so it seemed that an AY1119 would work well as a tripler. Success was immediate and so simple it took some time to convince myself that the output was on the right frequency.

The total collector-lead length of one inch is approximately resonant at 1267.5 MHz to provide maximum drive to the LO trough line. The trough-line portion of the converter is similar to that described in the ARRL vhf handbook and originally appeared in *QST,* March 1961.

The LO injection signal is coupled to the diode together with the signal to produce the desired i-f output on 28.5 MHz. A neutralized Motorola MPF-107 jfet is used in the i-f preamp to provide the lowest possible noise figure. An MPF-102 could be used for this stage, but it's more difficult to neutralize due to the higher feedback capacitance and may not produce as good a noise figure as the MPF-107. As this type of diode mixer has a considerable conversion loss, the noise




**Hole A to clear 0.001-uF feedthroughs**

**Hole B to clear 5-pF trimmers**

**Hole C to clear L1**

**Hole D to clear C13 and C14 tuning screws**

**Hole E tapping holes for bottom cover**

**Hole F 3/8" diameter**

**Hole G clearance hole for Q5**

**Hole H clearance for Q2 — Q4 (approx. 7/32" square)**

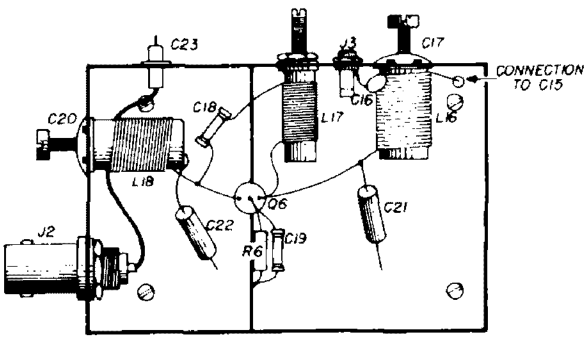
**fig. 2. Parts layout for mixer, oscillator, and multiplier chain. Arrangement should be followed as closely as possible. Q5 collector lead length is critical; it should not exceed 1 inch.**

figure of the i-f preamp contributes directly to the overall noise figure of the converter. If a 14-MHz i-f is chosen, then the MPF-102 would probably be suitable, but some degradation of overall noise figure may result due to poor image rejection.

## construction

The general layout and dimensions are given in **figs. 2** through **5**.

Some variation may be required to use components on hand. This shouldn't be a problem as long as all leads are kept as short as possible. The LO chain was constructed separately, then soldered to




**Hole A to clear 0.001-uF feedthrough**

**Hole B to clear 5-pf trimmer**

**Hole C 3/8" diameter**

**Hole D to clear J3**

**Hole E to clear L17**

**fig. 3. Preamplifier parts layout. Circuit is mounted in a separate box, which is secured to the top of the main chassis with self-tapping screws.**

the main chassis after adjustment (see below). This was convenient for the prototype, but the chassis could be constructed from one piece, if desired, with suitable partitions. I originally intended to construct the i-f amplifier in the compartment at the end of the mixer diode, but this would have made the diode inaccessible, so the preamp was constructed in a separate box and secured to the top of the converter with ¼ inch no. 2 self-tapping screws.

Type 2 BA or no. 10-32 countersunk screws are used for tuning screws at the center of each trough line. This size provides a fine thread for tuning, with a large diameter that reduces wobble. A nut is soldered to the top of the chassis. The end of the tuning screw can be slotted before threading into position. Both half-wave lines of ¼-inch O D copper tubing are soldered centrally in the trough lines after the tuning screws have been fitted. The signal-input loop of no. 18 wire is soldered to the connector, threaded through the mounting hole, out through a small clearance hole in the end plate, then soldered into position after the connector is tightened.

## mixer assembly

The mixer diode mount is constructed from tinplate as shown in **fig. 4.** A 3/4 x 3/16-inch strip is cut almost through at intervals of 1/16 inch to form fingers. The strip is then bent around a ¼-inch drill to fit the diode body. The seam is soldered, then the base of this section is soldered to the capacitor plate, C15. Remove all burrs and form the fingers to provide a firm fit on the diode body.

The capacitor is formed by a thin layer of teflon or polyethylene between C15 and partition 2 (**fig. 4**). The mounting screws land inside the ends of the two ¼-inch copper tubes, L11 and L12. The heads of the screws are insulated from C15 plate with small washers. The diode pin contact may be salvaged from an old bakelite octal wafer socket or may be fashioned from a small piece of tinplate. Solder a length of no. 18 tinned copper wire to the contact (L14), bend as shown in **fig. 2**, and solder to partition 2 in the signal trough line. It's not advisable to use a good diode while soldering, as it could be damaged by heat. Assemble the diode mount, C15, and check for shorts before inserting the diode.

## local-oscillator chain

Construction of the LO chain on the L-section shown in **fig. 4** is straightforward. The holes in the partition shields for Q2-Q5 should be a neat fit. Bend the emitter and collector leads at right angles before insertion, but take care not to

rotate the leads. If the angle is incorrect, straighten the lead and rebend in the desired direction. The base lead is soldered hard up to the transistor case. top of the converter with four ¼-inch no. 2 self-tapping screws. The lid is a press fit. A baseplate is desirable to reduce radiation from the trough lines.

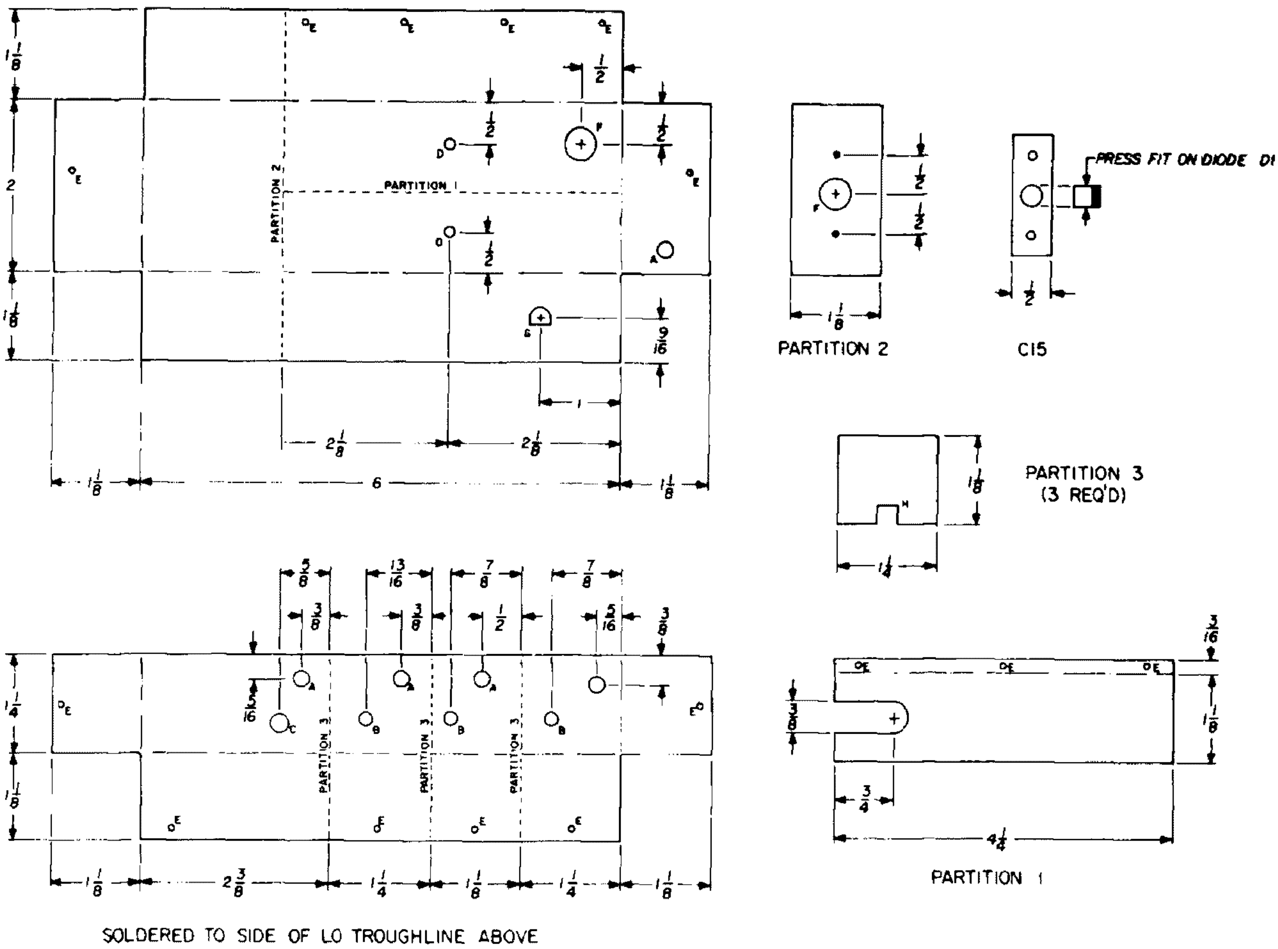



**fig. 4. Main chassis dimensions. Material is 26-gauge tinplate, which provides good shielding and mechanical stability. C15 is insulated from partition 2.**

This is important, as base lead inductance degrades the performance of the stage. There is no room for a heat sink, but this isn't necessary as the manufacturer's data sheet states, "soldering temperature not to exceed 300° C for more than ten seconds."

Tin the chassis first, then use a hot iron as quickly as possible. I've removed and replaced one transistor several times with no detectable reduction in performance. Once the multiplier chain is operating satisfactorily, solder this section to the side of the trough line and install Q5.

The i-f preamp is constructed in a simple box **(fig. 5)** and attached to the

## adjustment

You'll need some simple test equipment and a couple of easily made accessories for adjusting and aligning the converter circuits:

1. Signal generator.
2. Grid-dip oscillator.
3. Multimeter.
4. General-coverage receiver.
5. Diode mixer.
6. Signal-source termination.

The last two items are easy to make from junk-box parts. The diode mixer

(fig. 6) is used for checking crystal-oscillator performance. Almost any diode will work, but greater sensitivity will be obtained with a detector diode or a line. If a suitable gdo is available, each circuit may be tuned before wiring the 12-V supply line. Connect a multimeter (0-10 mA range) from C7 to the chassis.

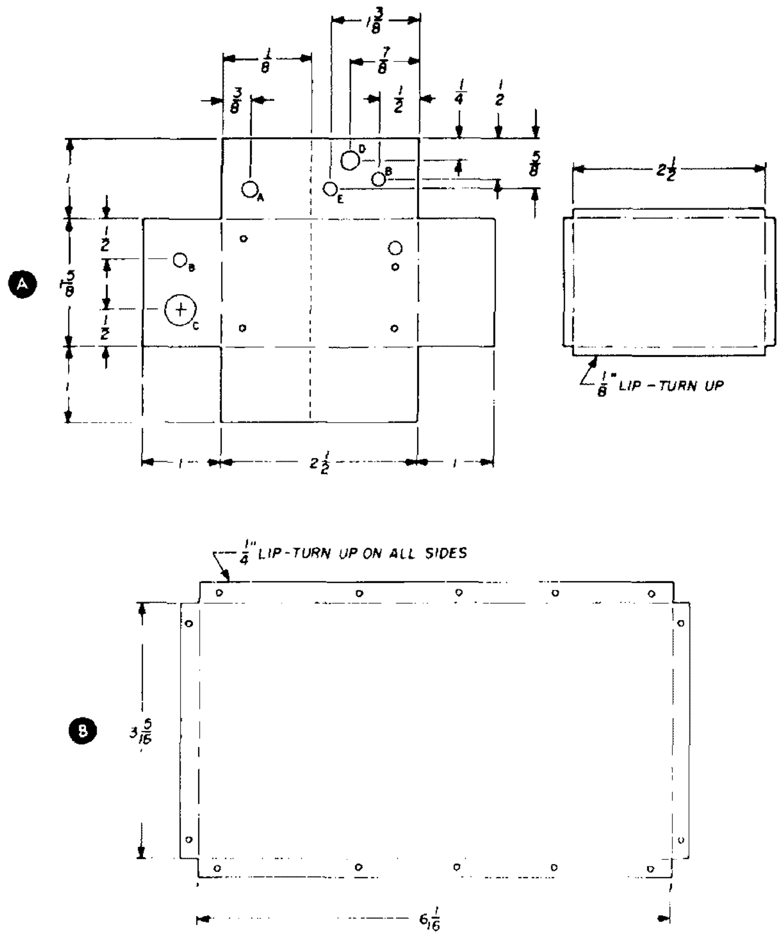



**fig. 5. Dimensions for preamp chassis, A, and the main chassis bottom plate.**

high-speed computer diode. The diode mixer should be enclosed in a small shielded box, and good-quality coax cable should be used for the coupling line.

The signal-source termination (fig. 7) is used in the procedure for optimizing the converter mixer noise figure. Mount the two connectors on a U-shaped bracket. Use either 75- or 50-ohm resistors, depending on the cable to be used. Make leads as short as possible. Use only carbon-composition resistors, as the spiral-track type are very reactive above 30 MHz or so.

## LO-chain adjustment

The LO chain is most conveniently adjusted before soldering to the trough

Slowly bring the gdo up to L4 until a reading of 1-2 mA is obtained on the meter. Tune the gdo for maximum current, taking care not to exceed full scale. The current peak indicates the resonant frequency of L4.

Trimmer C4 should now be adjusted so that the first doubler stage resonates at 105.625 MHz. The turns spacing of L4 may require adjustment if resonance occurs outside the range of C4. Pretune the remaining multiplier stages similarly.

This adjustment method has several advantages. Monitoring the collector current of Q2 ensures that transistor ratings won't be exceeded, especially when using a tube-type gdo with high output. Second, as the application of

power to the transistor changes circuit resonant frequency, compensation may be made by operating the collector at the approximate current to be used in the circuit. Third, many gdo's exhibit a very poor dip, particularly on the higher ranges. This method is not subject to false dips and provides good sensitivity.

## crystal oscillator

Connect the 12-V supply to C6 and connect C7 to the supply via a 0-10 mA meter. Adjust crystal-oscillator tuning for maximum current, which should be about 4 mA. L2 should be coupled as loosely as possible consistent with oscillator starting. If the coupling is too tight, the oscillator may revert to fundamental operation or even run free.

It's unlikely that a receiver covering this range will be available to check these conditions. This can be resolved by using your station receiver and a signal generator. The signal generator, which should be set to about 0.1=V output, is fed into the simple diode mixer (fig. 6). A one-turn link on the end of a length of coax is used to couple the oscillator signal. The output of the mixer is fed to the receiver tuned, for example, to 14.0 MHz. (Any frequency clear of stray pickup may be used.) The difference between, say, the third-overtone frequency, 52.8125, and 14.0 MHz is 38.8125 MHz. Some signal generators may not operate above 30 MHz, so the second harmonic of 19.406 MHz may be used. When the signal generator is tuned to the correct frequency, a strong beat note should be heard in the receiver with the bfo on. Carefully check for any spurious oscillations for at least ±1 MHz. A beat note produced by the crystal oscillator may be confirmed by detuning L1 slightly; or, alternatively, sufficient frequency shift usually occurs if the hand is brought close to L1.

If spurious oscillations are found, it may be necessary to decrease the value of R3 and recheck the coupling of L2. This coupling must be as loose as possible, consistent with reliable oscillator starting, when L1 is tuned slightly to the "slow" side of the peak. When the oscillator is operating correctly, there should be no output on the fundamental frequency (17.604 MHz). Check this by tuning the generator to 3.604 or 31.604 MHz and searching for a beat. The latter frequency is the more desirable, as there is less chance of it being confused with a harmonic from the signal generator, This method may also be used in reverse to check the calibration of a signal generator at several points with known crystal oscillators.

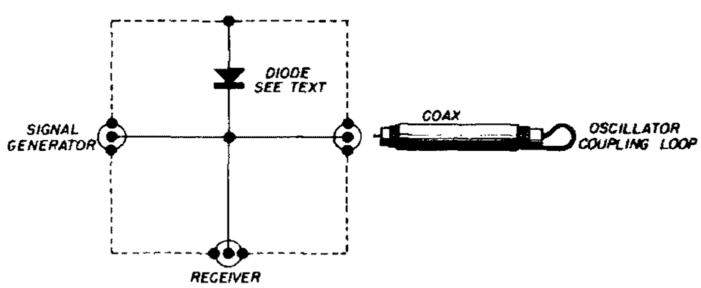



**fig. 6. Diode mixer accessory for checking oscillator performance.**

## multipliers

When the oscillator is operating satisfactorily, adjust L3 coupling until Q2 collector current is 4-5 mA. L2 and L3 should now be secured into position to prevent any movement.

Connect the 0-10 mA meter from C9 to the 12-V supply, and tune C4 for maximum Q3 collector current. It may be necessary to adjust the turns of L4 for the peak to occur near the center of the range of C4. Adjust L5 coupling to produce 6-7 mA of Q3 collector current.

Proceed with the adjustment of L8 and L9 similarly to give approximately 7 mA of collector current in Q4. It's unlikely that C8 or C10 have sufficient range to tune to the wrong harmonic.

The oscillator section should now be soldered to the trough line portion of the converter. Mount Q5 with the collector lead as shown in **fig. 2**. The overall length of the lead to the top of the feed-through capacitor should be about one inch. Apply power and peak C10 for maximum Q5 collector current. Adjust link L9 for 2.5-3 mA Q5 collector current, rechecking C10 tuning. Mount the i-f preamp and connect. Tune the LO cavity screw for maximum mixer diode current, which should be 0.5-1 mA. If the tuning

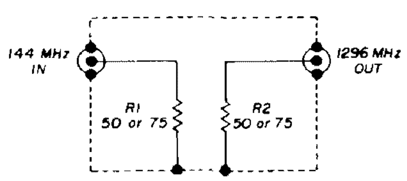



**fig. 7. Signal-source termination used for noise-figure optimization adjustments. Resistors are carbon composition types.**

of the early LO stages is checked, the tuning may appear very broad because of multiplier saturation. It's safer to check each individual stage collector current, except that L8 and C13 may be tuned for maximum mixer current.

The operating frequencies of L8 and L11 may be checked with Lecher lines by observing a dip in collector current when the lines are link coupled to the appropriate collector tuned circuit. (Lecher lines are described in most handbooks.) If the trough-line peak occurs with a gap of about 1/16 inch for C13, all multipliers are probably operating correctly. This completes the LO chain adjustment.

## i-f preamp

Apply power and if necessary adjust the value of R6 to give 4-5=mA drain current. Connect the output of the preamplifier to the receiver, which should be tuned to 28.5 MHz. Normally the stage will oscillate over a considerable portion of L17's range. Adjust L17 until the oscillation ceases, then tune to the center of the "stable area." Peak L16 and L18 for maximum noise in the receiver, and recheck L17. It may help to link couple an external signal to peak the input and output circuits. Due to large variation in fets, it may be necessary to add or remove turns from L17. Final adjustments should be made for best noise figure.

## checking mixer noise figure

Most amateurs don't have access to a good noise generator; therefore a weak 1296-MHz signal is necessary to optimize mixer noise figure. The harmonic of a 144- or 432-MHz transmitter will suffice. The resistive termination (**fig. 7**) is used for this check. Reduce transmitter output to about ¼ watt and connect the transmitter to load resistor R1 of the termination.

If a mixer diode such as the 1N21 or 1N23 is used, it should now be possible to detect a harmonic from a 432-MHz transmitter connected via the terminating unit. The type number of most shf mixer diodes is followed by a letter; e. g., 1N23F. The higher the letter, the lower the noise figure – and also the higher the price. As usual, a compromise is required unless a diode is obtainable free!

Mixer noise figure is best optimized with a signal near the noise level. Connect a low-range ac voltmeter or vtvm across the receiver output. It may be necessary to couple directly across the output transformer via a capacitor to obtain sufficient noise level for a reading on the voltmeter of, say, 0.5 volt. It's not necessary to remove the agc if the signal is kept very low.

Apply the signal to the converter, and tune for maximum indication on the meter. If the indication is more than about 1 V, it will be necessary to reduce transmitter power output or decrease the coupling between R1 and R2 on the terminating unit. If the signal level is much higher, it will be difficult to detect the small changes that indicate if one is proceeding in the right direction.

## final adjustments

Tune the receiver a few kHz off the signal and, if necessary, adjust the receiver gain control to give the reference 0.5-V noise-level reading. Retune to the signal and note the signal level. Make an adjustment and note the difference between noise and signal level. As some adjustments affect the overall gain, it will be necessary to make small adjustments to the receiver gain control for 0.5-V reference noise level before noting the signal level. (We are looking for an increase in signal over noise.) When this ratio exceeds 2:1, reduce the signal level slightly and continue. This may sound tedious, but it can be performed quite rapidly with practice.

The mixer trough line may be peaked initially by tuning until a dip is noted in diode current then screwing C14 out slightly, which tunes this circuit higher in frequency.

The adjustments controlling the noise figure are:

**1. Signal trough.** Normally tuned for maximum signal.

**2. Mixer current.** Alter injection in increments of 50 $\mu$A to find the optimum level, which is normally 0.2-0.3 mA, but will depend on the diode. The injection level may be conveniently controlled initially by detuning C10. Once the optimum level is found, the coupling of L9 may be adjusted to give this value with L8 peaked.

**3. Diode coupling.** The area enclosed by the link should be close to that shown in fig. 2. Try altering the area by varying lead length in 1/8-inch steps. Once again, this will depend on the diode.

**4. Input coupling.** The area of the link controls matching and should be close to that shown.

**5. I-f preamp.** The adjustment of L17 and the input coupling of L15 are critical for best noise figure. Adjust L17 in small steps, repeak C17 and C20, and check signal-to-noise ratio. The number of turns on the coupling link, L15, should also be varied.

After optimizing these adjustments, I measured the converter noise figure on a commercial noise generator. It was 9.8 dB, which appears to be about as good as can be expected with a simple mixer using this type of diode.

## power feed

Note that the bottom of L19 is connected to the 12-V line. This was a simple expedient to feed power to the converter via the coaxial i-f cable, thus allowing the converter to be mounted close to the antenna. I used a modified BC-454 command receiver converted to 28-30 MHz with link coupling to the input of the rf stage. The bottom of the link was returned to the 12-V supply line in the receiver. No degradation of overall noise figure or gain resulted. Also, the problem of a separate battery feed was eliminated. My 144-MHz converter was likewise adapted, so that changing from 144 to 1296 MHz required changing only the i-f cable, which is convenient for portable work.

If you don't wish to modify the i-f receiver, an isolating capacitor and choke may be used to feed the 12 V into the coax. If this feature is not required in the converter, return the bottom of L19 to the chassis in the usual manner.

## conclusion

The construction and adjustment of a simple but effective 1296-MHz converter has been described in detail in the hope that its simplicity may encourage some of the dc boys to "have a go." No special test equipment is required and, with the exception of the mixer diode, the set uses inexpensive and readily available components. VK4KE constructed a similar converter using silver-plated brass and obtained almost identical results. There appears to be little advantage in silver plating other than for appearance. Time permitting, I'll describe the construction of the varactor triplers and antennas used on this project.

**ham radio**

# how to use the smith chart

Jim Fisk, W1DTY

**Although articles on the Smith chart** have appeared in amateur magazines from time to time, amateurs have made little use of this handy transmission-line calculator – probably because it has been difficult to measure complex impedances with simple homebuilt equipment. However, this problem has been solved with the simple impedance bridge described by W2CTK – at least for the high-frequency range.[1] With careful attention to lead dress and component layout his instrument should be usable on six and two meters.

A hasty glance at the Smith chart suggests a formidable array of curved lines and circles that would cause the most hardened technician to go into fits of despair. On the other hand, if you spend a little time with the chart and look at each of its component parts, it's not really very complicated. Perhaps the one thing that scares many prospective users is its unfamiliar circular shape; it's not at all like the straight-line graphs you're accustomed to. However, when you understand the chart and have mastered its use you'll be able to solve complex impedance and transmission-line problems much easier and faster then ever before.

## layout of the chart

The Smith chart is basically a circle which contains various circular scales. The horizontal line through the center marked "resistance component" is the only straight line on the chart and is called the "axis of reals" (see **fig. 1**). Constant resistance circles are centered on the axis of reals, tangent to the rim of the chart at the infinite resistance point.

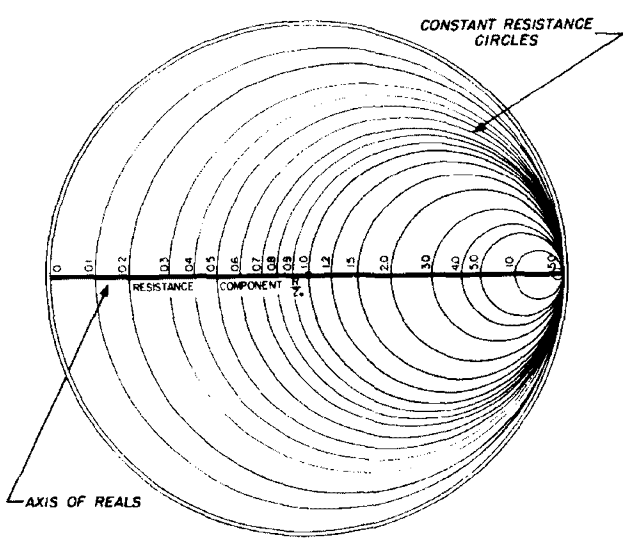



**fig. 1. Smith chart resistance scales.**

All the points along a constant-resistance circle have the same resistive value as the point where it crosses the axis of reals.

Superimposed upon the resistance-circle pattern are portions of other circles tangent to the axis of reals at the infinite

resistance point, but centered off the edge of the chart (fig. 2). The large outer rim of the chart is calibrated in relative reactance and is called the "reactance axis." Any point along the same constant-reactance circle has the same reactive value as the point where it intersects the reactance axis on the rim of the chart. All points on the Smith chart above the axis of reals contain an inductive-reactive component and those below the axis of reals contain a capacitive-reactive component. Since the calibration points go from zero to infinity, *any* complex impedance can be plotted on the chart.

The impedance coordinates on the Smith chart would be of little use without the accompanying peripheral scales (fig. 3). These scales relate to quantities which change with position along a transmission line. Two scales are calibrated in terms of wavelength along the transmission line: one, in a clockwise direction, is "wavelengths toward generator," and the other, counter-clockwise, is "wavelengths toward load." The entire length of the circumference of the chart represents one-half wavelength.

## normalized numbers

Normalized values must be used when plotting impedances on the Smith chart.*

Normalized impedance is defined as the actual impedance divided by the characteristic impedance of the transmission line.

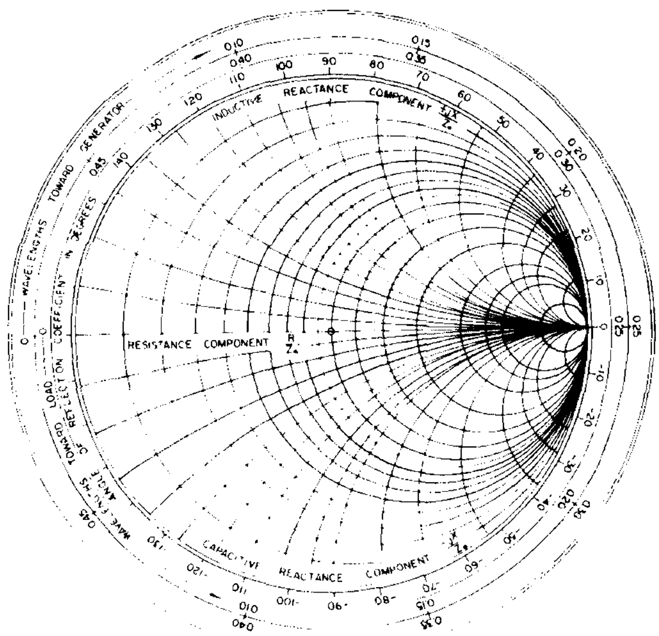



**fig. 3. Smith chart peripheral scales.**

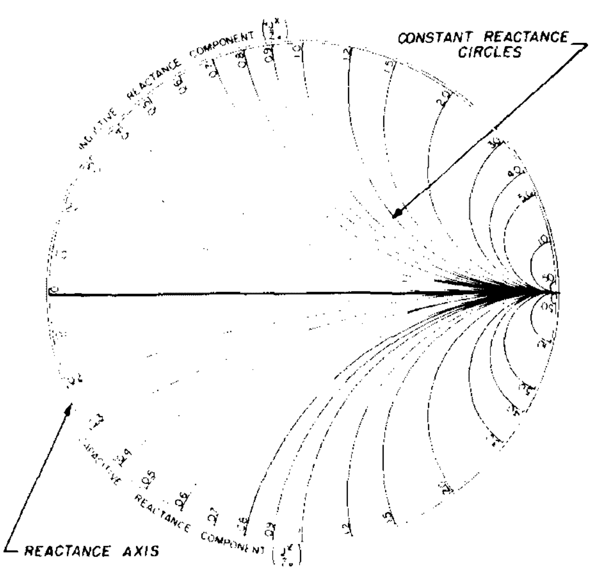



**fig. 2. Smith chart reactance scales.**

Normalizing is done to make the chart applicable to transmission lines of any and all possible values of characteristic impedance. For example, a 50-ohm coaxial transmission has a normalized value of 50/50 or 1. On this basis an impedance of 120 ohms would have a normalized value of 120/50 = 2.4 ohms. Similarly, $Z' = 0.8$ ohms (the prime indicates a normalized value) would correspond to a value of 0.8 times the characteristic impedance of the line or 0.8 X 50 = 40 ohms.

What has been said about coaxial cable with regard to normalized impedance applies equally to waveguide, where a *characteristic impedance of 400 ohms at* a specific frequency would be considered unity in normalized form. All other values would be related to this value, so

*Since 50-ohm systems are standard for military and industrial use, 50-ohm Smith charts are available. On a 50-ohm Smith chart the center point has a value of 50 ohms.

that a 560 - ohm component would have the value 560/400 = 1.4 ohms in normalized terminolgy, while $Z'$ = 0.9 in normalized form would actually be 0.9 X 400 = 360 ohms.

## plotting values on the chart

Any complex impedance, regardless of value, may be plotted on the Smith chart. For example, assume the load on a 50-ohm transmission line is 42.5 -j31.5 ohms. This is equal to 0.85 - j0.63 when normalized. To plot this point on the chart, locate 0.85 on the axis of reals and note the corresponding constant-resistance circle (fig. 4). Next locate 0.63 on the periphery of the chart. The quantity (-j) indicates a capacitive-reactive component so the value 0.63 is on the lower half of the chart. Note the constant-reactance circle representing -j0.63. The complex impedance 0.85 -j0.63 is at the intersection of the constant-resistance and constant-reactance circles.

Draw a line from the center of the chart through this point to the outer rim. With the point 1.0 on the axis of reals as the center, scribe a circle that intersects the impedance point. This circle is known as the "constant-gamma circle," and its radius is equal to the coefficient of reflection. The constant-gamma circle crosses the axis of reals at two points; the point of intersection to the right of center is the standing wave ratio (2.0 in this case).

If the voltage were measured at this point on the transmission line, it would be found to be at a maximum. Conversely, the point of intersection one-quarter wavelength away on the left-hand axis of reals is a point of voltage minimum (this point is also equal mathematically to the reciprocal of the swr).

The point at the intersection of the radial line and the angle of reflection coefficient scale represents the phase of the coefficient of reflection. This is the angle by which the reflected wave leads or lags the incident wave. When these two waves add in phase to give maximum voltage, the impedance is resistive and greater than the characteristic impedance of the line and the angle of the coefficient of reflection is zero. As you move away from the zero-phase-angle point in a clockwise direction toward the generator the reflected voltage lags the incident voltage, and the phase angle is negative for the first quarter wavelength. The reactive component of the impedance in this region is negative or capacitive.

At the quarter-wavelength (90°) point the incident and reflected waves are out of phase and the angle of the coefficient of reflection is ±180°. As you continue in a clockwise direction the two waves become increasingly more in phase and between one-quarter and one-half wavelength from the voltage maximum the reactive component is inductive, the reflected wave leads the incident wave, and the reflection coefficient has a positive angle.

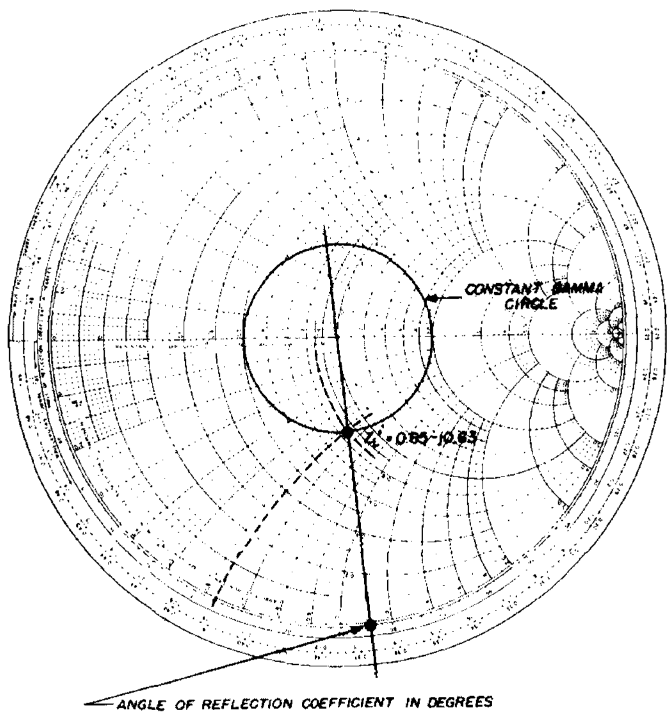



**fig. 4. Plotting impedance coordinates on the Smith chart.**

A number of parameters are uniquely related to one another as well as to the magnitude of reflections from the load and are conveniently plotted as scales at the bottom of the Smith chart. These

parameters are vswr, coefficient of reflection, vswr in dB, reflection loss in dB and attenuation in 1-dB steps.

## using the smith chart

The general utility of the Smith chart is best illustrated by showing examples of its more common uses. Use of the radially-scaled parameters will be shown in the same way.

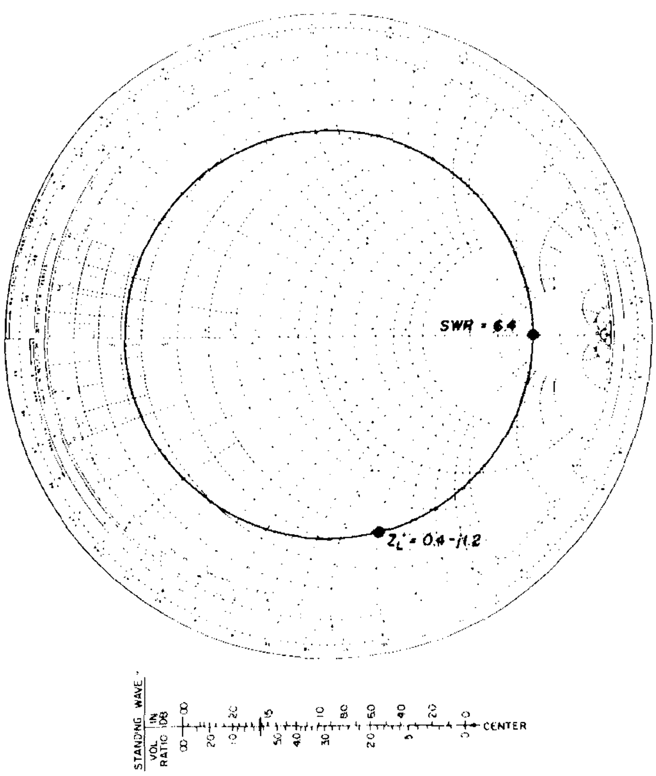



**fig. 5. Using the Smith chart to find swr (example 1).**

**example 1. Finding standing-wave ratio.**

A 75-ohm transmission line is terminated with a load impedance $Z_L$ = 30 - j90 ohms. What is the swr? (See **fig. 5.**)

**1.** Normalize the load impedance by dividing by 75

$$\frac{30 - j90}{75} = 0.4 - j1.2$$

**2.** Locate this point on the chart.

**3.** Construct a constant-gamma circle so its circumference passes through this point.

**4.** The swr is defined by the point where the constant-gamma circle crosses the axis of reals on the right-hand side. In this case swr = 6.4.

**5.** The swr may also be determined with the radial nomograph. This is simply accomplished by marking a distance equal to the radius of the constant-gamma circle on the radial scale labeled "standing wave voltage ratio." The value of swr in dB may also be determined from this scale.

$$swr_{dB} = 16.1 \text{ dB}$$

**example 2. Finding the reflection coefficient ($\Gamma$) and angle of the reflection coefficient ($\alpha$) for voltage and current.**

A 50-ohm transmission line is terminated with a load impedance 65 – j75 ohms. What is the reflection coefficient and angle of reflection coefficient? (See **fig. 6.**)

**1.** Normalize the load impedance

$$\frac{65 - j75}{50} = 1.3 - j1.5$$

**2.** Locate this point on the chart and draw a line from the center of the chart through it to the outer scale.

**3.** Construct a constant-gamma circle.

**4.** The reflection coefficient may be calculated by measuring the radii of the constant-gamma circle and the Smith chart to its first periphery and by computing their ratio. Smith-chart radius = 57/16 inch; constant-gamma radius = 32/16 inch.

$$\Gamma = \frac{32}{16} \div \frac{57}{16} = 0.56$$

**5.** The coefficient of reflection may also be found on the radial nomo-

graph. Simply mark the radius of the constant-gamma circle on the scale labeled "reflection coefficient of voltage." The constant-gamma radius intersects the radial scale at 0.56. The "reflection coefficient of power" may also be determined from this same scale at 0.314.

6. The angle of the reflection coefficient is defined by the intersection of the radial line plotted in **step 2** and the "angle of reflection coefficient in degrees" scale on the rim of the chart.

$$a = -46^\circ$$

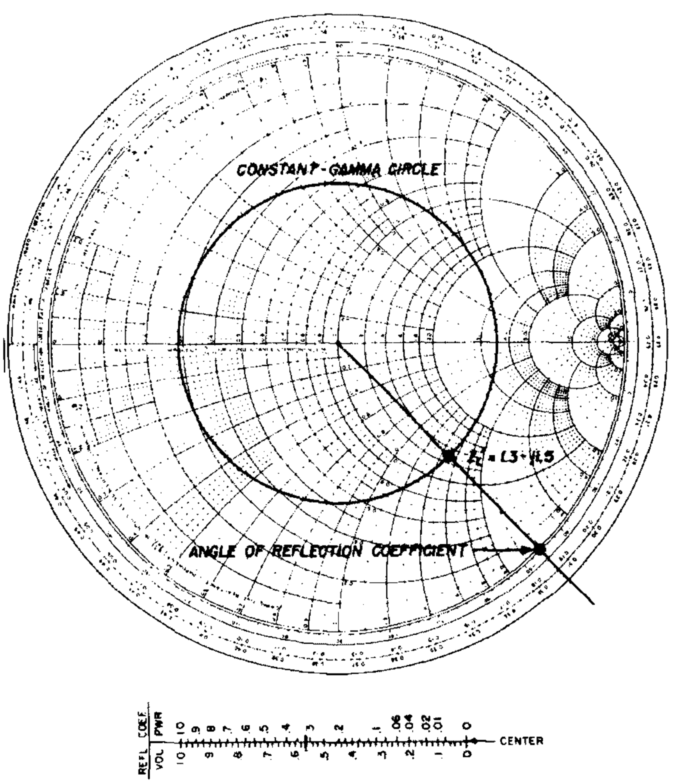
**example 3. Finding input impedance.**

A 50-ohm transmission line 20 feet long is terminated with $Z_L = 50 - j50$ ohms. What is the input impedance at the sending end of the line at 14.1 MHz? (See **fig. 7.**)

**1.** Normalize the load impedance

$$\frac{50 - j50}{50} = 1 - j1$$

**fig. 6. Finding reflection coefficient with the Smith chart (example 2).**




**2.** Find the length of the transmission line in meters by multiplying by 0.3048.*

20 feet x .3048 = 6.096 meters

**3.** Find the electrical length of the transmission line at 14.1 MHz. First, determine the wavelength at 14.1 MHz. Free-space wavelength is found

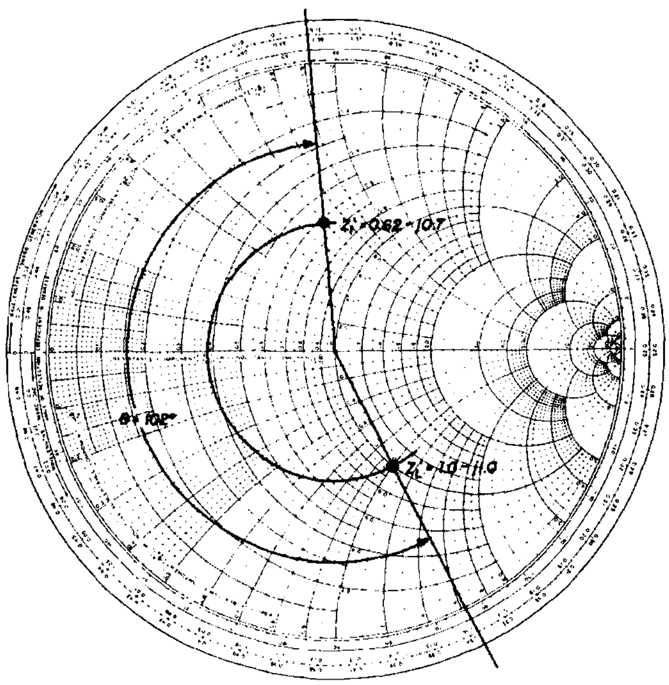


**fig. 7. Using the Smith chart to find input imepedance (example 3).**

by dividing the speed of light by frequency

$$\lambda = \frac{3 \times 10^8 \text{ meters per second}}{14.1 \times 10^6 \text{ cycles per second}} = 21.276 \text{ m}$$

Calculate the electrical length of the transmission line

$$\theta = 360^\circ \left(\frac{6.096 \text{ m}}{21.276 \text{ m}}\right) = 102^\circ = 0.28 \text{ wavelength}$$

**4.** Plot the impedance coordinates from **step 1** on the chart and draw a line from the center of the chart through this point to the outer scale.

*Although all the computations may be made in feet (or inches) the metric equivalents are somewhat easier to work with. To convert from inches to centimeters, multiply by 2.54.

5. Draw another line from the chart center to the outer scale at a point 0.28 wavelength clockwise (toward the generator) from the line drawn in **step 3.** Swing an arc from the center of the chart through $Z_L'$ to this line. The intersection is at $Z_L' = 0.62 + j0.7$, the *normalized input impedance.* To find the actual impedance this value must

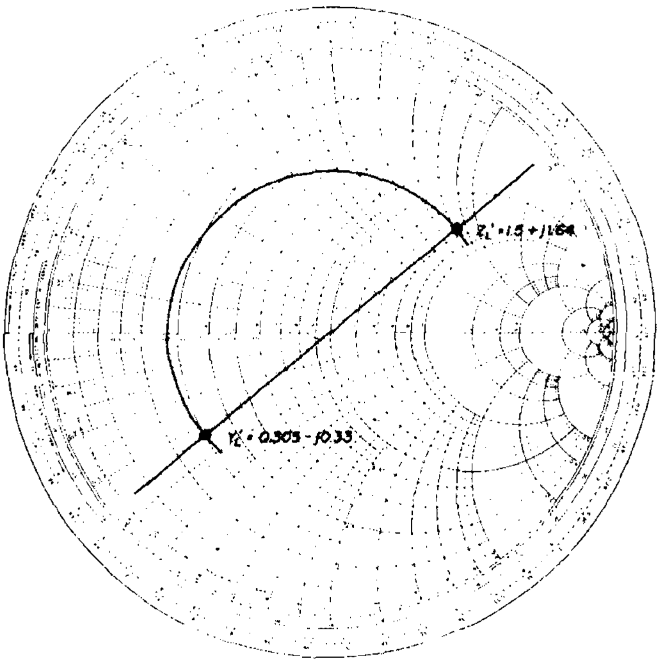



**fig. 8. Calculating load admittance (example 4).**

be multiplied by the line's characteristic impedance

$$Z_i = 50(0.62 + j0.7) = 31 + j35$$

**example 4. Calculating load admittance.**

The impedance of a load terminating a 50-ohm transmission line is 75 + j82 ohms. What is the admittance of the load? (See **fig. 8.**)

**1.** Normalize the load impedance

$$Z_L' = (75 + j82)/50 = 1.5 + j1.64$$

**2.** Plot this point and draw a line through the center to the outer scale on the oppsoite side of the chart.

**3.** Swing an arc through $Z_L'$ to the line on the opposite side of the chart. The point of intersection denotes the *normalized* admittance

$$Y_L' = 0.305 - j0.33$$

**4.** Calculate the actual admittance by multiplying the characteristic admittance of the system times the normalized admittance. The characteristic admittance ($Y_O$) is equal to the reciprocal of the characteristic impedance

$$Y_O = \frac{1}{Z_O} = \frac{1}{50} = 0.02 \text{ mho}$$

Therefore, the admittance is

$$Y_L = 0.02(0.305 - j0.33)$$
$$= .0061 - .0066 \text{ mho}$$

**example 5. Determining the effect of a characteristic impedance change.**

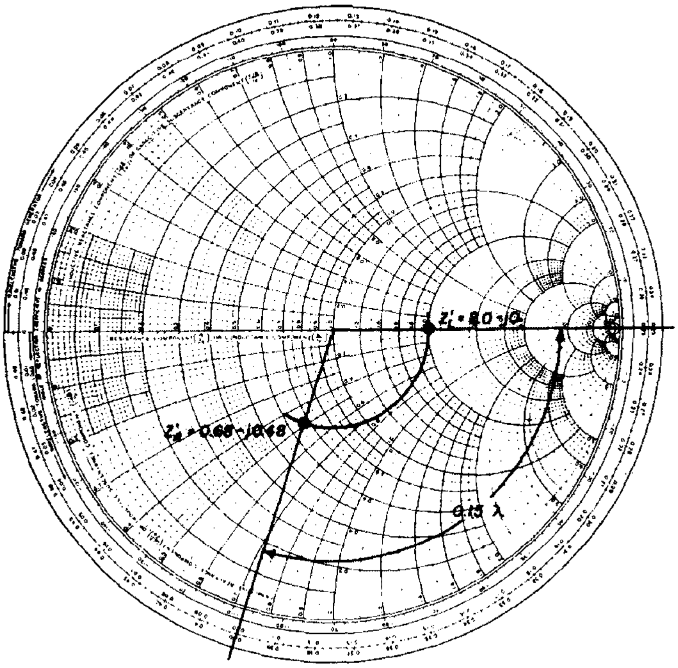
A 50-ohm transmission line, 0.15 wavelength long, is terminated with 100 - j0 ohms. The 50-ohm line is fed from a 72-ohm line. What is the vswr in the 72-ohm line? (See **fig. 9.**)

**1.** Normalize the load impedance

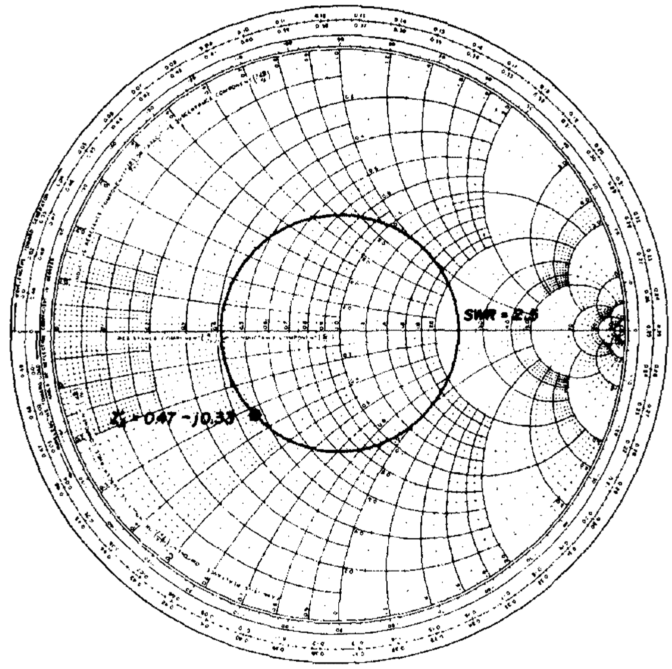
$$Z_L' = (100 - j0)/50 = 2 - j0$$

**2.** Determine the input impedance at the point where the two transmission lines are connected, 0.15 wavelength from the load. Plot the normalized load impedance on the chart and draw a line from the center of the chart through this point. Note that the line crosses the "wavelengths toward generator" scale at the 0.25 wavelength mark (**fig. 9A**).

**3.** Move 0.15 wavelength in a clockwise direction along the "wavelengths toward generator" scale to the 0.40 wavelength mark. Draw a line from this mark through the center of the chart. Swing an arc through $Z_L'$. The intersection of the arc and the radial line denote the input impedance to the




A




B

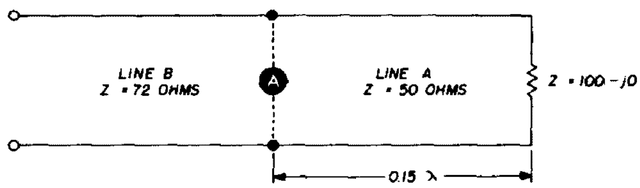
50-ohm transmission line 0.15 wavelength from the load

$$Z_A' = 0.68 - j0.48$$

4. Find the impedance at point A (**fig. 9C**) and normalize to the 72-ohm line. The impedance at point A is 50(0.68 - j0.48) = 34 - j24 ohms. Normalize this value to the 72-ohm line

$$(34 - j24)/72 = 0.47 - j0.33$$

5. Plot this point on the chart (**fig. 9B**) and draw a circle through $Z_A$ to the "axis of reals." The vswr in the 72-ohm line is 2.5:1. The vswr can also be found with the radial nomograph as outlined in **example 1.**




C

**fig. 9. Determining the effect of a characteristic imepedance change (example 5).**

In the upper vhf region ordinary capacitors and inductors cannot be relied upon to act as pure reactances, and sections of transmission line are often used in their place since *any* input reactance may be obtained with the proper length of open- or short-circuited line.

**example 6. Transmission lines as circuit elements.**

It is desired to obtain +j100 ohms reactance with a 50-ohm short-circuited

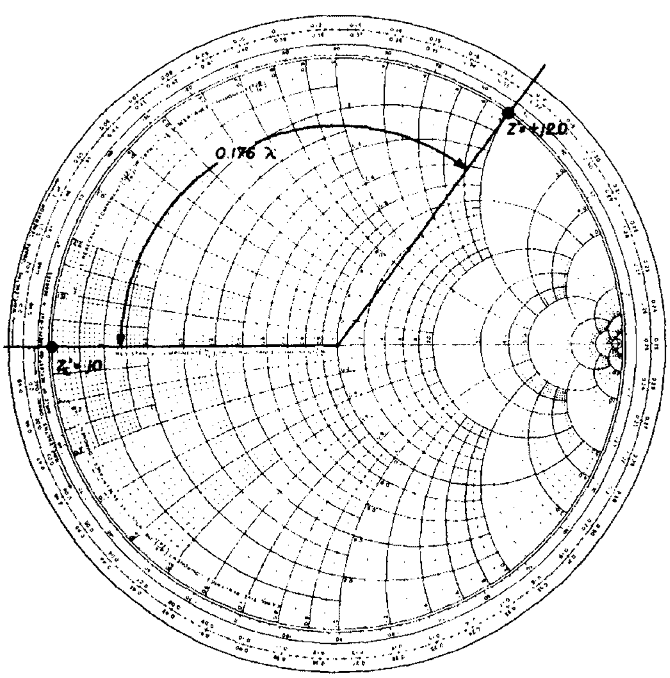



**fig. 10. Using a transmission line as a circuit element (example 6).**

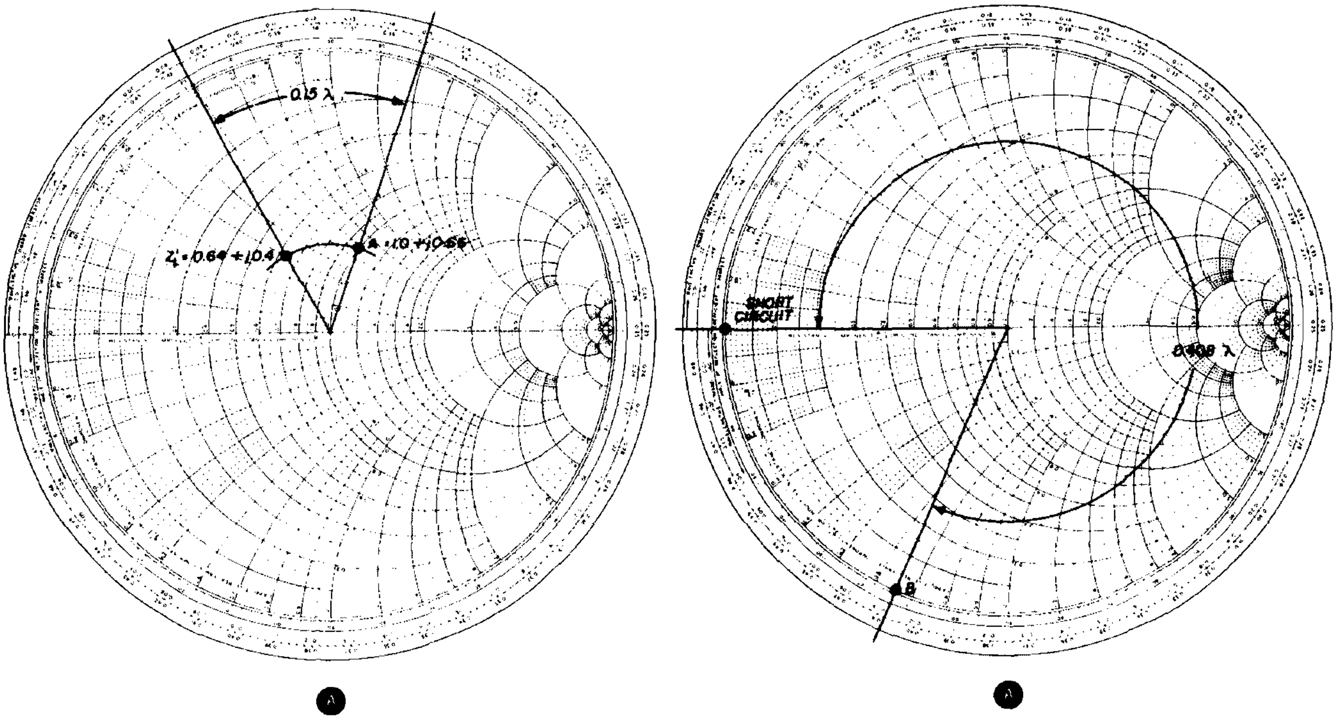



**fig. 11. Finding matching stub length and location (example 7).**

transmission line as the circuit element. What length is required? (See **fig. 10.**)

**1.** Normalize the desired reactance

$$Z' = (+j100)/50 = +j2$$

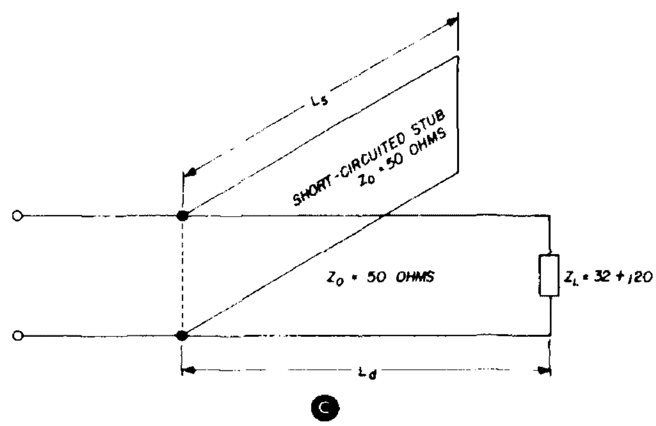
**2.** Since the line is short-circuited,

$$Z_L = 0 + j0, \text{ and } Z_L' = 0 \text{ ohms.}$$

**3.** Plot these two points on the chart and draw lines from the center of the chart through each of them. On the "wavelengths toward generator" scale there is a distance of 0.176 wavelength between the two lines. Therefore, a transmission line 0.176 wavelength long is required for a reactance of +j100. (At 144 MHz, +j100 represents an inductance of 0.11 $\mu$H.)

**example 7. Finding matching stub length and location.**

A 50-ohm transmission line is terminated with a load impedance of 32 + j20 ohms. A matching stub is to be used to provide a match to the line. Both the length of the stub ($L_s$) and its distance from the load ($L_d$) are variable; find $L_s$ and $L_d$. (See **fig. 11.**)




**1.** Normalize the load impedance

$$Z_L' = (32 + j20)/50 = 0.64 + j0.4$$

**2.** Locate this point on the chart and draw a line through it to the outer scale. The line crosses the "wavelengths toward generator" scale at the 0.085 mark.

**3.** Construct a constant-gamma circle through the impedance point, terminating it at the unity resistance circle (point A in **fig. 11B**).

4. Draw a line through point A to the outer scale of the chart. This line crosses the "wavelengths toward generator" scale at 0.15. $L_d$, the position of the matching stub, is the distance between the two lines on the "wavelengths toward generator" scale.

$$L_d = (0.15 - 0.085) = 0.065\ \lambda$$

5. To find the length of the stub, determine the amount of reactance necessary to match out the load. The required reactance is the difference between the reactance at point A and the reactance at the center of the chart. The reactance at point A is +j0.66; the reactance at the center of the chart is +j0. The required stub reactance is

$$j0 - j0.66 = j0.66$$

6. Locate the reactance −j0.66 on the rim of the chart (point B, fig. 11C). Determine the distance between the short-circuit point and the required reactance (point B) along the "wavelengths toward generator" scale. $L_s$ = 0.408 wavelength.

This matching technique can be used on the high frequencies as well as vhf. On 15 meters for example (21.3 MHz), using 50-ohm coax, a 158-inch stub would be placed 25 inches from the load. On 432 MHz a 7.82-inch stub would be placed 1.24 inches from the load.

## lossy lines

All the examples shown so far have assumed no attenuation in the transmission line. Since all lines have some loss, this must be considered to find the actual case. However, at many amateur frequencies loss is low enough to be neglected. Nevertheless, at 144 MHz and above, line loss should be considered when using the Smith chart.

Attenuation along a uniform transmission line causes the impedance point to spiral inward toward the center of the chart when moving toward the generator; when moving toward the load the impedance point spirals outward toward the rim of the chart. The rate at which the spiral approaches the center (or the rim) depends upon the attenuation as well as the starting point. Impedance points near the rim are affected more per dB of attenuation than points near the center.

The attenuation effect is easily determined with the scale at the bottom of the Smith chart labeled "transmission loss, 1-dB steps." Since the initial point on this scale must apply to any point on the chart, it is laid out without numerical calibration. The opposite attenuation effects of moving toward the load as opposed to moving toward the generator are indicated by arrows on the scale which show the proper direction to move the corrected impedance point. Thus, to determine the effect of 2-dB attenuation, simply mark off two 1-dB intervals in the proper direction along the scale from the initial starting point before reading the actual impedance coordinates.

**example 8. Impedance transformation through a lossy line.**

A 50-ohm transmission line 24 centimeters long is terminated with 10 − j10 ohms. What is the input impedance to the line at 250 MHz if the attenuation of the line is 2 dB? (See fig. 12.)

1. Normalize the load impedance

$$Z_L' = (10 - j10)/50 = 0.2 - j0.2$$

2. Find the electrical length of the line at 250 MHz.

$$\lambda = \frac{300 \times 10^8}{250 \times 10^6} = 120 \text{ cm}$$

The electrical length of the line is

$$\theta = 360^\circ \left(\frac{24\text{cm}}{120\text{cm}}\right) = 72^\circ = 0.2 \text{ wavelength}$$

3. Plot the impedance from step 1 on the chart and draw a line through this point to the outer scale.

4. Draw another line from the chart center to the outer scale at a point 0.2 wavelength clockwise (toward the generator) from the line passing through $Z_L'$. Swing an arc through $Z_L'$ to this line. The intersection point denotes $Z_i'$ = 0.71 + j1.52 ohms. This is the normalized solution for the lossless case. The rf energy from the generator is *attenuated* 2.0 dB on reaching $Z_L$, and the voltage reflection coefficient is 2.0 dB lower than the lossless case.

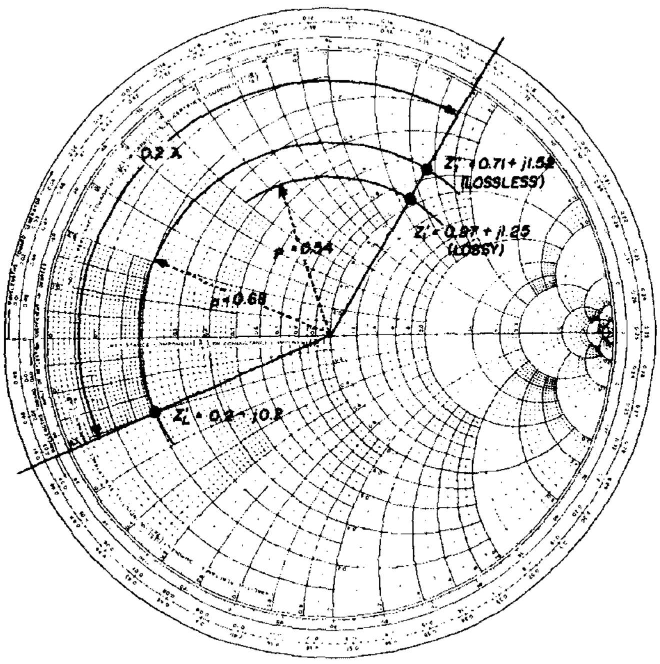



**fig. 12. Impedance transformation through a lossy transmission line (example 8).**

5. The reflection coefficient ($\rho_O$) for the lossless case is 0.68 (found on the scale at the bottom of the chart). The actual reflection coefficient ($\rho_1$) is 2.0 dB below $\rho_O$. Since 2.0 dB represents 0.794 voltage ratio, the actual coefficient of reflection may be calculated by multiplying the lossless coefficient of reflection by this ratio

$$0.794\rho_O = 0.794\ (0.68) = 0.54$$

6. Swing an arc equal to the ratio $\rho_1$ = 0.54 so it intersects the line drawn through $Z_i'$; the radius of this arc can be found on the "voltage reflection coefficient" scale on the bottom of the chart. The normalized impedance for the lossy case is 0.97 = j1.25. The actual input impedance is

$$Z_i = 50\ (0.97 + j125) = 48.5 + j62.5 \text{ ohms}$$

## slotted lines

At frequencies above 300 MHz conventional impedance-measuring instruments give way to the *slotted line.* A slotted line is essentially a section of transmission line with a small opening so you can use a probe to measure the voltage along the line. Vswr is easy to determine with the slotted line since it's the ratio of the maximum voltage along the line to the minimum. With the known vswr and position of the first voltage minimum, it's easy to find the impedance of the load with the Smith chart.

**example 9. Calculate the load impedance from the vswr and position of the first voltage minimum.**

A 50-ohm transmission line has a vswr of 2.5; the first voltage minimum is 0.1 wavelength from the load. What is the impedance of the load? (See fig. 13.)

1. Draw a radial line from the center of the chart through the 0.1 wavelength mark on the "wavelengths toward load" scale.

2. *Find the 2.5 point on* the axis of reals and draw a constant-gamma circuit through this point to intersect with the 0.1-wavelength line.

3. Read the coordinates of this intersection to obtain the normalized impedance of the load

$$Z_L' = 0.56 - j0.57$$

$$Z_L = 50\ (0.56 - j0.57) = 28 - j28.5 \text{ ohms}$$

If you use twinlead or open-wire feed-

line this technique could be used to determine the impedance of your antenna. However, the voltage probe must be held a uniform distance away from the line for all measurements, and must not be so close that it disturbs the electric field around the conductors.

## expanded smith charts

The more closely an antenna is matched to a transmission line, the closer the impedance points are to the center of the Smith chart. In a well-designed system the impedance points may be so close to the center of the chart that it's difficult to work with them. When this happens it's best to use an expanded Smith chart. Two versions are commonly available: one with a maximum swr of 1.59, the other with a maximum swr of 1.12.

The use of the expanded Smith chart is shown in **fig. 14.** In **fig. 14A** the impedance plot of a well-matched 10-meter beam over the low end of the phone band falls very close to the center of the chart. When these same impedance points are plotted on the expanded Smith chart in **fig. 14B** they are much easier to read and work with.

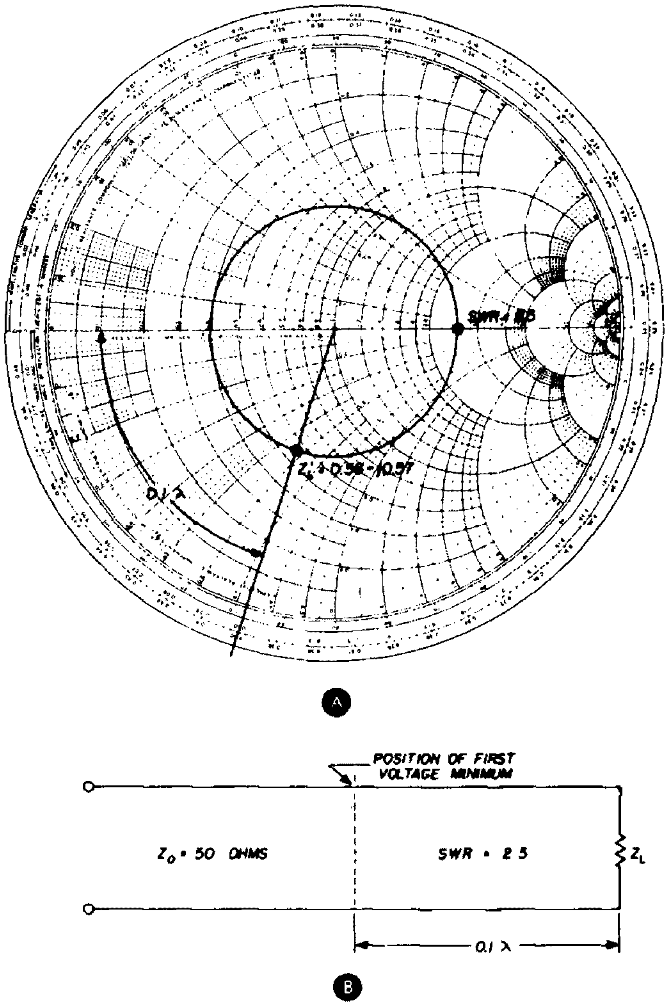
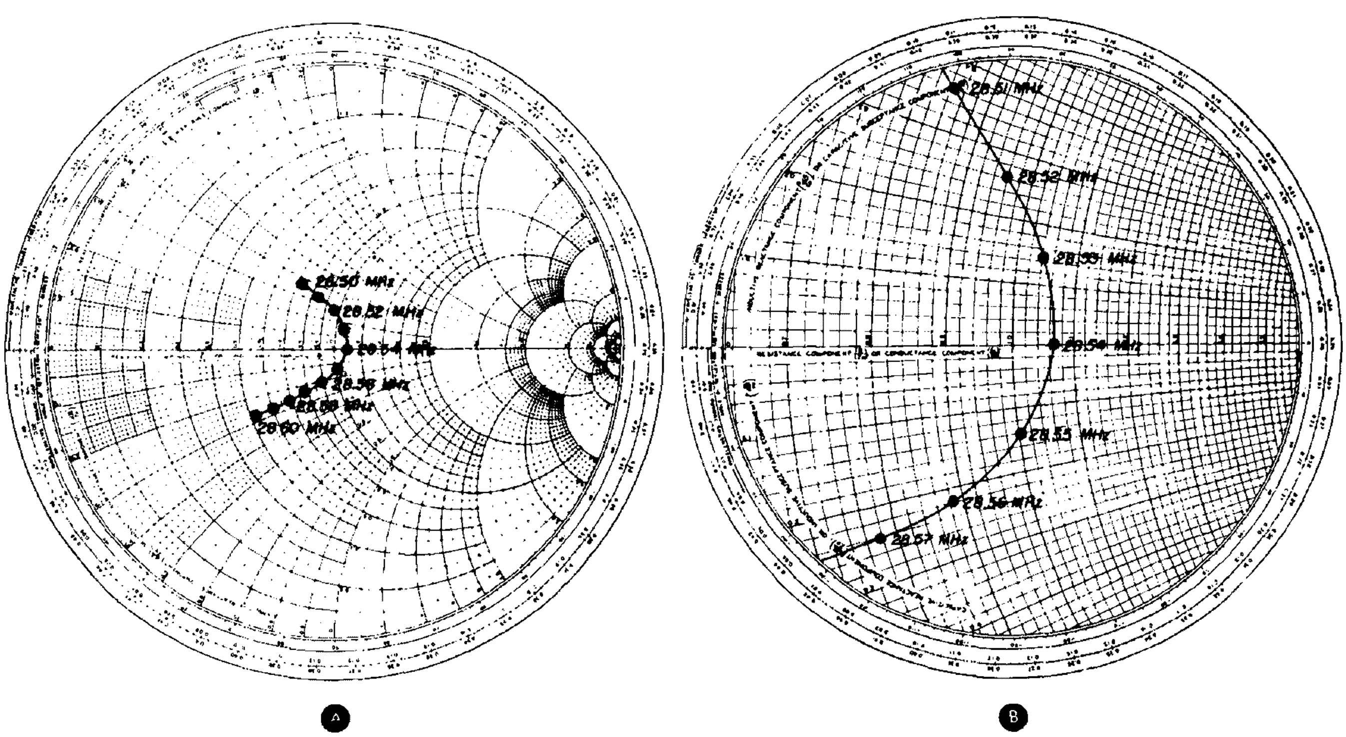



**fig. 13. Using the Smith chart to find load impedance from vswr and position of the first voltage minimum on a slotted line (example 9).**

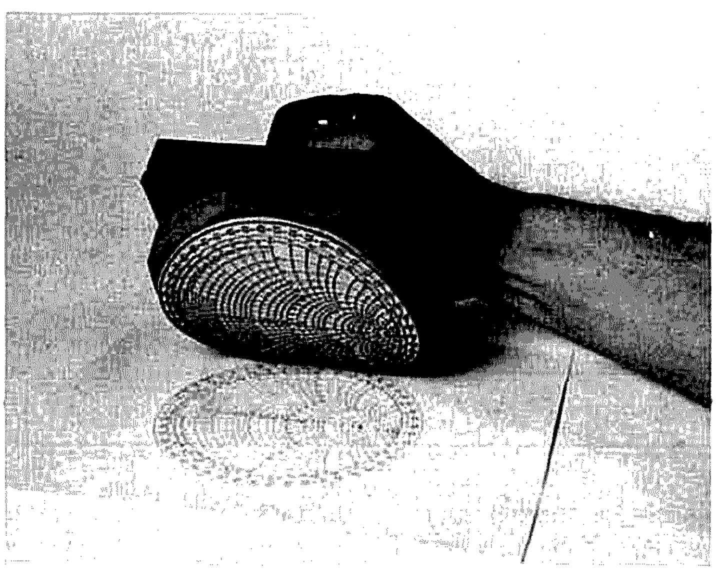
**fig. 14. Use of the expanded Smith chart. Impedance points in A are too close together; expanded chart in B is easier to work with.**



## where to buy them

Smith charts can usually be purchased at college bookstores in small quantities, or in larger quantities from Analog Instruments Company or General Radio.* If you buy directly from the manufacturer the minimum quantity is a little large so it might be a good idea to get your club to sponsor the purchase.

Another solution is the Smith-chart rubber stamp shown in the photo. This stamp is 10 cm (about 4 inches) in diameter and presents an adequately detailed grid structure for most engineering problems. The rubber surface of these stamps is cast from metal dies, and is dimensionally compensated for rocker-mount ellipticity and shrinkage. The capacity is well over a million impressions so you should never be able to wear it out. The stamps are available in standard (vswr = ∞) or expanded form (vswr = 1.59 or 1.12) from the Analog Instruments Company. Cost is $14.75 each.



**Smith chart rubber stamp is 10 cm in diameter.**

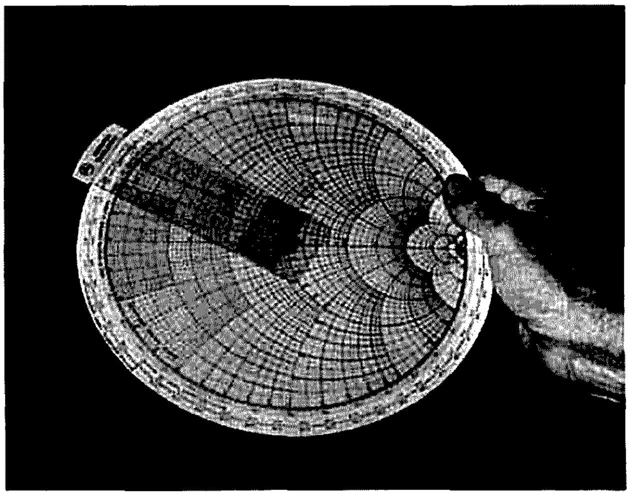


**fig. 15. Smith chart calculator provides rapid answers to complex impedance problems.**

If you don't need a permanent record of your Smith chart calculations, the calculator shown in fig. 15 provides rapid answers to complex impedance problems. This calculator is constructed from two laminated plastic discs and a radial arm pivoted at the center with a sliding cursor. A circular slide rule is provided on the reverse side. Complete instructions are furnished. Priced at $3.00 from Amphenol RF Division, 33 E. Franklin Street, Danbury, Connecticut 06810; ask for the Amphenol RF Calculator.

*Smith charts from Analog Instruments come in packages of 100 sheets, $4.75 the package. For standard charts order 82-BSPR; expanded charts (maximum swr = 1.59), order 82-SPR; highly expanded (maximum swr = 1.13), order 82-ASPR. Analog Instruments Company, Post Office Box 808, New Providence, New Jersey 07974

Smith charts from General Radio are available in pads of 50 sheets, $2.00 per pad. For standard charts, normalized coordinates, order 5301-7560; 50-ohm coordinates, order 5301-7569; normalized, expanded coordinates, order 5301-7561. General Radio, West Concord, Massachusetts 01781.

references


1. Henry S. Keen, W2CTK, "A Simple Bridge for Antenna Measurements," *ham radio*, September, 1970, p. 34.
2. Phillip H. Smith, "Transmission Line Calculator," *Electronics*, January, 1939, pp. 29-31; and "An Improved Transmission Line Calculator," *Electronics*, January, 1944, pp. 130-133 and 318-325.
3. Phillip H. Smith, "Electronic Applications of the Smith Chart in Waveguide, Circuit and Component Analysis," McGraw-Hill, New York, 1969.


ham radio

# injection laser experiments

A progress report on further development of pulsed-light circuits for one-way ranging

Ralph W. Campbell, W4KAE, 316 Mariemont Drive, Lexington, Kentucky 40505

In an earlier article I described my experiments with light-emitting diodes operating in the near infrared.[1] Pulsed power supplies, photodetectors, and optical enhancement devices were also discussed. These early experiments were a first approach toward understanding the more sophisticated injection laser for use in a communications link.

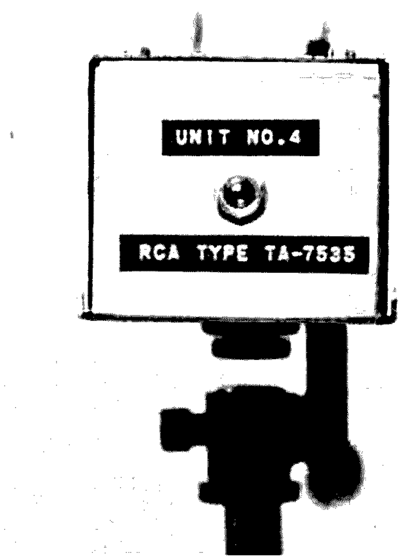
This article describes the results of my work with injection lasers and related devices. I hope it will provide an incentive for further experimentation by amateurs who are interested in exploring new frontiers in electronics.

## pulsed-light devices

After almost a year of research, I learned that injection lasers were in their infancy—much like the bipolar transistors of 20 years ago. On the development market a 2-watt (peak) injection laser costs $36-$50. RCA's TA-2628 gallium arsenide injection laser (3 watts peak) is listed in their Fall 1969 pricing sheet.* It

*The TA-2628 is "inventory limited" according to information obtained from an RCA field office. Their May 15, 1970 listing shows 166 pieces in stock at $36.25. Their recommended replacement is TA-7606; same price. editor

**Injection laser/pulser power supply. Sample 4 of RCA's developmental TA-7535 is used. Total power output is 1W minimum.**



is recommended as the least-expensive injection laser diode for experimental work. Also recommended is the pulsed power-supply circuit included in the TA-2628 data sheet.

As mentioned in my previous article, the limiting element in an LED communications system is the detector. This is equally true for systems using injection lasers. Although other photodetectors are mentioned as substitutes and replacements in the following circuits, the type PIN-10 broad-area Schottky photodiode, made by United Detector Technology,* is the recommended device for serious work. This is the detector with which I achieved a communication range of more than 1000 feet with my experimental injection laser equipment. I found that it was impossible to detect ambient noncoherent light sources with the PIN-10, so you won't have to worry about street lights overloading the detector.

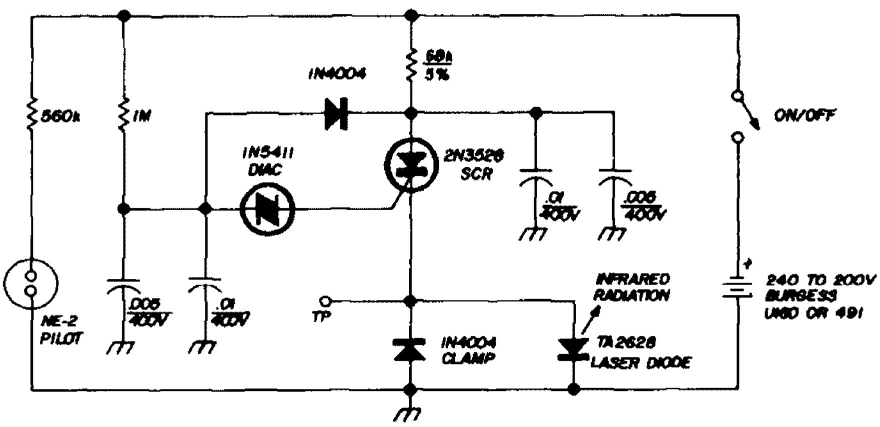



**fig. 1. A 30-ampere (peak) pulsed power supply for the RCA TA-2628 injection laser. This circuit is recommended for experimental cut-and-try work. (Courtesy RCA.)**

## laser ranging

For amateur work, cw or keyed pulse (like A2 emission) is the best modulation method for use with injection lasers. I expect to see a 2-mile range in as many years using a 10-watt-peak injection laser at 9000 angstroms. Estimated keyed-pulse communication ranges at the present time, based on my research, are shown in **table 1**. Transmitting optics must be used in all cases. Further range enhancement would require an S1 photomultiplier with a filter and huge fresnel optics. Average power is $10^{-3}$ to 5 x $10^{-3}$ of the peak values shown in the table.

My PIN-10 detector was operated in the back-biased photoconductive mode. With a special network, it can be used as a photovoltaic detector. I will be happy to furnish this information to serious experimenters upon receipt of a self-addressed stamped envelope.

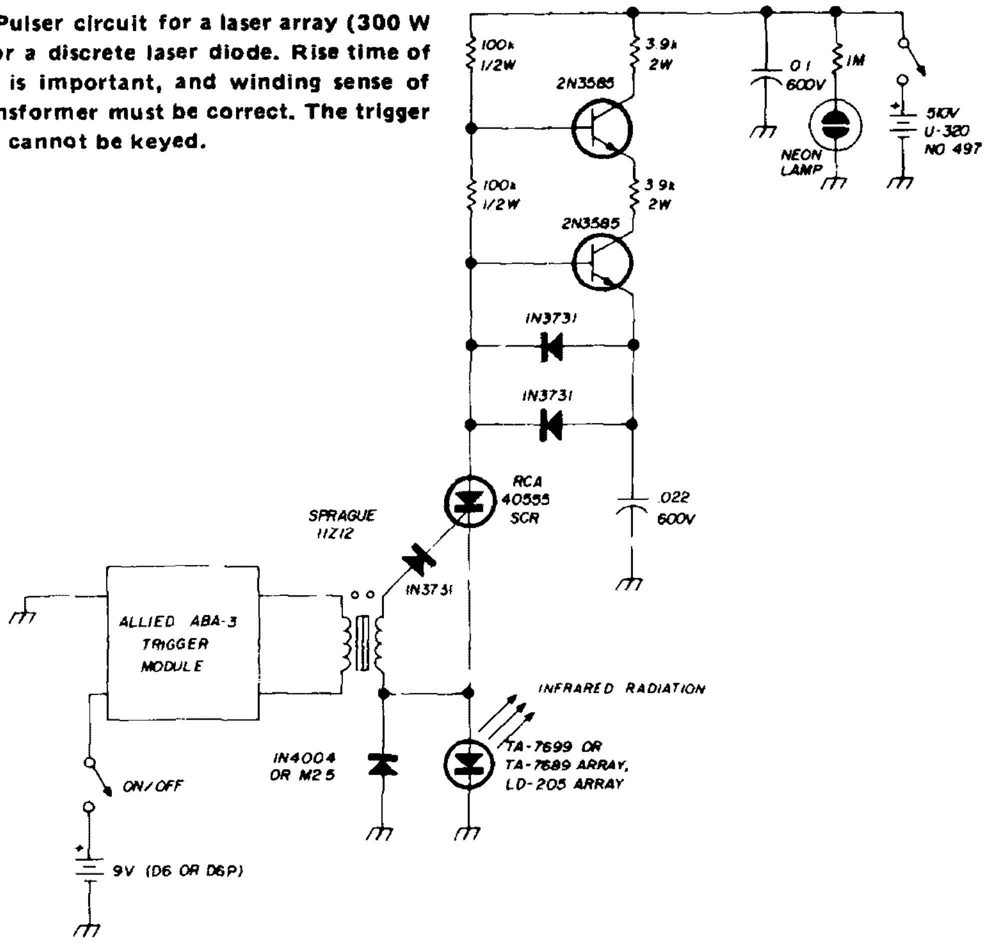
True photoconductive operation should not be confused with majority-carrier photoresistive detectors, such as the cadmium selenide cells. These devides are sluggish performers when used with pulsed light at pulse repetition rates above 500 Hz. Their only advantage is high sensitivity, which is a worthwhile tradeoff in certain applications.

## detector parameters

A handy aid for determining the operating characteristics of a silicon photodiode is the "Silicon Photo Detector Calculator" made by UDT. This is a plastic slide-rule-type device similar to reactance calculators and the like. It allows one to design a thermal-agitation, noise-limited, back-biased network for a specific application. The UDT calculator was indispensable for this project.

*United Detector Technology, 1732 21 St,. Santa Monica Calif. 90404.

fig. 2. Pulser circuit for a laser array (300 W peak) or a discrete laser diode. Rise time of the scr is important, and winding sense of the transformer must be correct. The trigger module cannot be keyed.




## sample calculation

The PIN-10 has the following parameters:

| | |
|---|---|
| dark current | $5 \times 10^{-7}$ A |
| junction capacitance | 100 pF |
| responsivity | 500 mA/W |
| minimum load resistance | 5k ohms (Johnson-noise limited) |

Using the calculator, we find:

| | |
|---|---|
| rise and decay time | 1.1 µsec |
| half-power cutoff freq. | 300 kHz |
| noise equivalent power | 3.6 µW (detector shot noise) |

My goal was to make the noise equivalent power (nep) of the 5k load equal to the shot noise of the biased detector. The actual nep is only $1.4 \times 10^{-13}$ watt with the 5k load resistor. To reconcile the difference, it's necessary to increase the value of the load resistor until a true noise match occurs. In this case, I used 500k. This match exists between the resistor shot noise and the detector shot noise. (Resistor noise is of the shot type when using a load resistor higher than 50k.)

The remaining problem is to determine the frequency response based on the loads. With a 500k load and a 100 pF junction capacitance, the half-power-point cutoff frequency is only 3 kHz. This is too low for applications with a prf of 500 Hz, because frequencies up to 5 kHz must be passed to retain the proper laser pulse shape. By using the lower scale of the calculator, it is found that 330k instead of 500k is the proper load resistance for photoconductive operation of

table 1. Estimated ranges for injection laser systems.

| GaAs laser power (watts peak) | bare PIN-10 detector (miles) | PEM detector* (miles) |
|---|---|---|
| 2 | 1 | 6 |
| 20 | 3.2 | 19.2 |
| 200 | — | 60 |
| 2 K | — | 192 |

*Developed by Santa Barbara Research Center

table 2. Summary of injection laser data from manufacturers' sheets.

| device type | peak power (watts) | peak reverse voltage (volts) | threshold current (amps) | forward current (amps) | pulse width (nsec) | rep rate (Hz) | duty cycle (%) | wavelength (A°) |
|---|---|---|---|---|---|---|---|---|
| TA-2628* | 3 | 3 | 12 | 30 | 200 | $20 \times 10^3$ | 0.02 | 9050 |
| TA-7535 | 2 | 2 | 4 | 10 | 200 | $5 \times 10^3$ | 0.1 | 9050 |
| LD205 array | 100 | 60 | 25 | 75 | 200 | 500 | — | 9000 |

Notes: 1. Maximum values at 25° C

2. Peak reverse voltage should not be exceeded

*replaced by TA-7606; same operating characteristic.

the PIN-10.

For photoresistive applications, such as shown in fig. 4, it's permissible to use a 33k load resistor. However, it should be borne in mind that a bipolar emitter-follower is used here, not an fet. If you'd like to experiment with a 50k low-noise carbon-film load resistor, you'll find that a Johnson/shot-noise match will occur with this input circuit. Back bias should be the same as shown (i.e., 10–15V).

I am now operating my PIN-10 with about 50V back bias into a 330k load resistor. The circuit is coupled to an MPF-107 jfet operating as a source follower. I've had excellent results with this circuit. A shot-noise match of both load resistor and the PIN-10 occurs at 5 kHz.

## radiation safety

Much has appeared in the press about the dangers of radiation from lasers. While it's true that a high-power pulsed ruby rod or glass laser presents a health hazard, this is not true of the low-power injection devices described in this article. Looking directly into the sun is much more dangerous than looking into an injection laser, even at what might be considered traumatic levels of 10 mW average power through collimating optics. Authorities on the subject have indicated that external average radiation levels of the order of 10 mW from a pulsed ruby rod are used for surgical work. An upper limit for such use is 2 W average power—not the 1–5 mW used with the injection lasers described here. So there's no reason for apprehension about radiation danger with the circuits shown.

## pulsed power supplies

The pulsed power supply shown in fig. 1 is based on the circuit in the RCA data sheet for the TA2628 GaAs injection laser. It evolved from long-distance phone conversations and much correspondence with RCA. I call it the "RCA Classic Pulser." It's the only reliable pulser suitable for extensive experimental work involving "cut, try, and innovate" designs.

The circuit is limited by the rise and fall times of the 1N5411 diac rather than

300W (peak) laser array pulser. This is the highest-powered laser-array pulser suitable for amateur work. Unit uses type LD-205, LD-205S, or TA-7692 array diodes. TA-7689 can be used with about 300V.

scr switching frequency. The 2N3528 scr has a T0-66 case, which is handy for cut-and-try techniques.

The 1N4004 diodes protect the laser and the scr. The 1-meg resistor in the diac circuit is preferable when using battery power, because the 2.2-meg resistor in the RCA data-sheet circuit for the TA-2628 resulted in no oscillation. Another replacement for the TA-2628 is the TA-7535, which I am now using. (See table 2.)

The circuit of fig. 2 is a pulser for a laser array such as the RCA TA-7689 or Laser Diode Laboratories' type LD-205.* The circuit can also be used to drive an RCA TA-7699 discrete injection-laser diode. The advantage of this circuit is that it provides a better match for an array load (up to 10 ohms impedance). If used with the TA-7699, the circuit will develop 3 to 5.5 times the necessary threshold current for low-impedance applications.

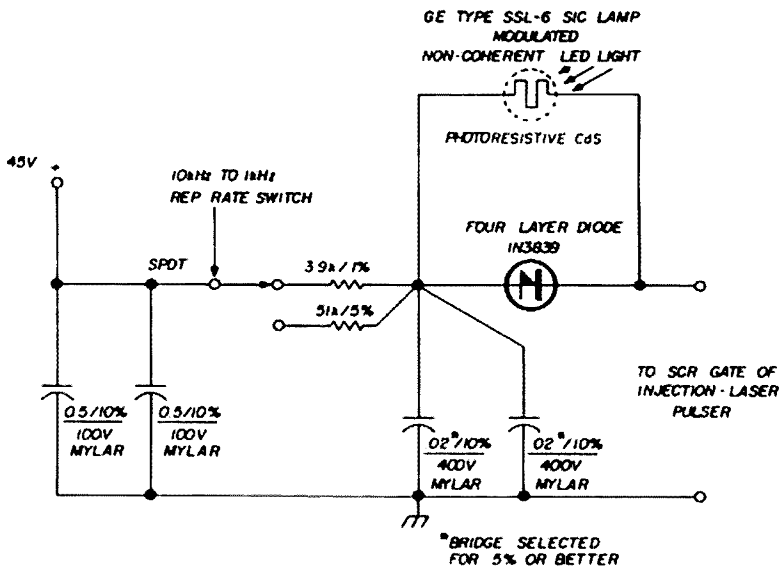
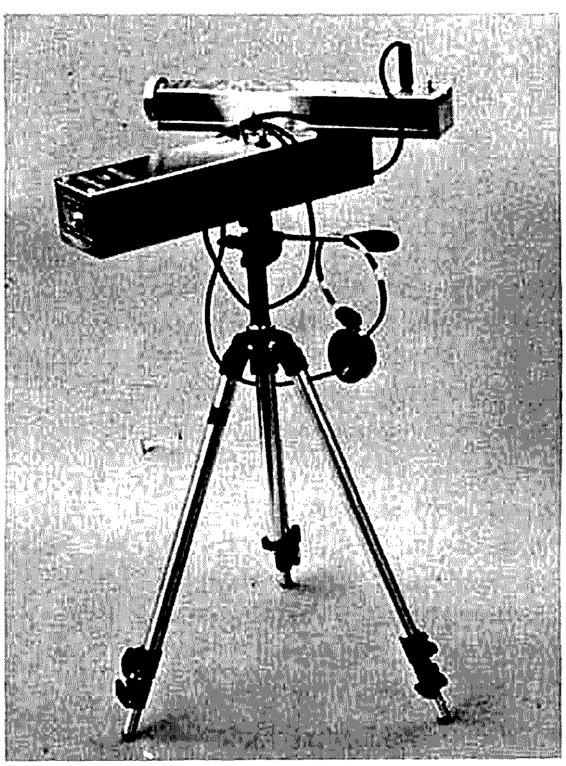



**fig. 3. One- or ten-kHz oscillator that can be frequency modulated for voice work. Typical turn-on and turn-off times of the 1N3829 are 20 and 40 nsec. (Courtesy RCA.)**

**Portable injection laser transmitter and detector. The detector (top) is the same unit described in author's article in ham radio on LED experiments using L14A502 photodiodes, which are replaceable with MRD-500's for best results.**



The external trigger circuit, using the Allied ABA-3 module and Sprague 11Z12 pulse transformer, allows random pulse trains to fire the scr. A possible application would be optical radar ranging.

Rise time of the scr is critical. Also, the winding sense of the transformer must be correct or nothing will happen. To-66 transistors were used, because they simplify the wiring (a must for short leads) and eliminate dependence on terminal strips.

*The source for RCA lasers is Solid-State Optical Engineering, RCA, U. S. Route 202, Somerville, N. J. 08876. Best source for laser pulsers and high-power arrays is Laser Diode Labs, 205 Forrest St., Metuchin, N. J. 08840.

## diode oscillator

The circuit of fig. 3 is strictly for experimental use. It is straight from RCA's research labs and is not one of my innovations. I am presenting it to show the state of the art as of this writing.


**The UDT type PIN-10 broad-area detector, recommended for laser diode receivers. The receiving optics were discarded, because surface pickup is important rather than narrow-beam optical gain with these detectors.**

An ITT four-layer diode, the 1N3839 ($7.73), can be used for oscillator service at 1 and 10 kHz. This device (and probably the APD4C50, made by American Power Devices, Inc.) are the only superfast triggers with timing response of less than 50 nsec that will allow injection-laser pulsing at the prf up to 10 kHz. The cadmium selenide cell in parallel with the 1N3839 can be used for voice work by frequency modulating noncoherent light input.

Before attempting to build and operate this circuit, I'd suggest that you write to me. I'll be glad to answer questions on design based on my latest data.

## detectors

As pointed out previously, the UDT PIN-10 is a must for serious work with injection-laser communications systems. The PIN-10 has broad-area response. Aspheric condensing lenses were used at first to gather received laser light from an aperture in space. It soon became obvious that the optics weren't necessary because of the broad-area response of the PIN-10. A better detector is the UDT PIN-25; however its cost may be beyond the resources of the experimental researcher.

A schematic of a bipolar laser detector is shown in fig. 4. This circuit was designed for CdSe cells, but will also work with so-called "exotic" devices such as the PIN-10. The 33k load resistor is optimum for an array of three cells. With the larger p-i-n diodes such as the PIN-25, bias would be −45V and the load resistor would be 1 meg. If you use, say, −9V reverse bias on a PIN-10, a 180k low-noise carbon-film resistor would make a satisfactory load.

A broad-area detector for use with type MRD500 photodiodes was developed, based on the circuit shown in fig. 9 of my previous article[1] on LED experiments. This modified circuit uses 7-10 MRD500s in an array that replaced the

**The M3 Sniperscope, battery, and high-voltage power supply. Unit is essential for collimating laser energy and boresighting.**

L14A502 phototransistors in the LED detector. Interested readers may obtain a copy of this broad-area detector circuit by sending me a self-addressed stamped envelope.

## construction hints

The photos show details for constructing the circuits. Improvisation is used for mounting large components. For example, the high-voltage dry cells were cemented in place. I used Deltabond thermal epoxy type 152 to mount the laser in an MX-1684/U coax termination, which had been drilled to 3/16-inch ID.

Problems with cementing lenses were resolved by mounting, checking beam tilt, returning the unit to an epoxy disintegrating solution, and repeating the work until a satisfactory mount was obtained. Oven curing was used to save time. A Bud MS-3050 minislide box chassis holds the detector in ranging tests. This Sniperscope was purchased for $50 from a surplus source. It was barely in working order, and I had to rebuild it. (One source quoted $325 for a Sniperscope, so it pays to shop around!)

The image tube in the Sniperscope is a type 6032. It has a peak-power conversion/sensitivity loss of 30 dB when used with pulsed devices at 0.1-percent duty cycle; thus it will not respond to peak power.

Since the instrument is basically an infrared telescope, it's difficult to use at distances greater than 100 meters or when the field of view is too narrow. About 450 feet is the practical limit from my experience.

Despite its limitations, the Sniperscope is essential for initial setup and adjustment of equipment before starting extended laser range tests. Instruments

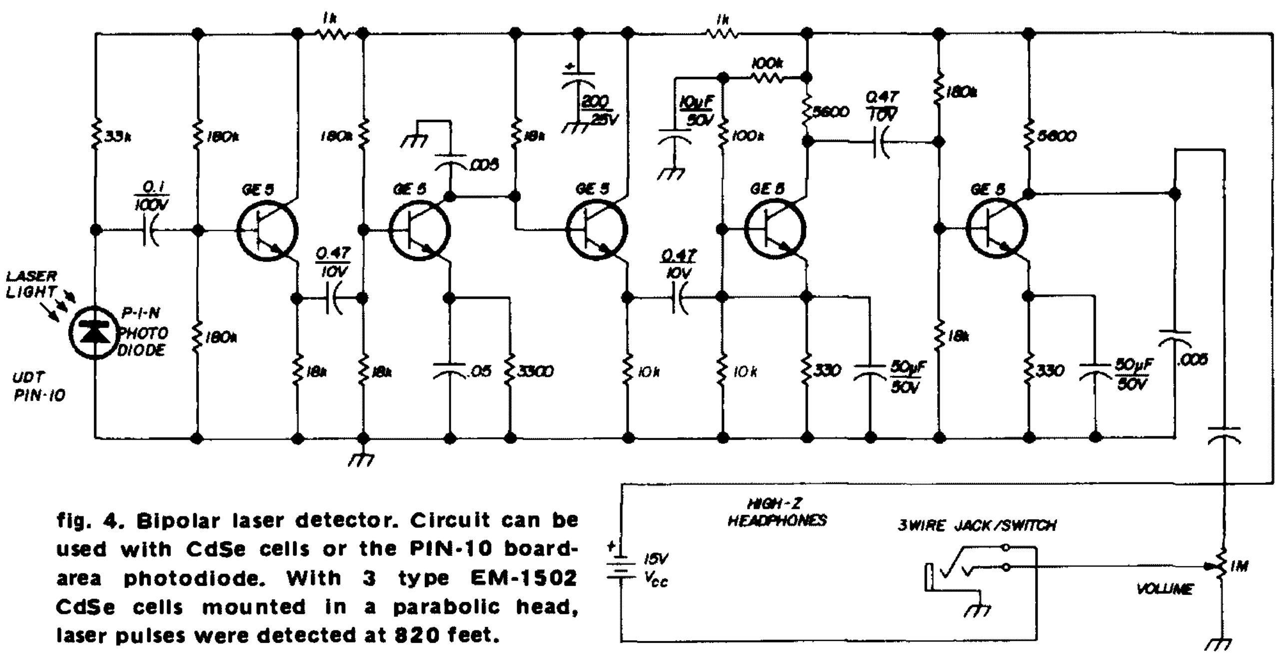



**fig. 4. Bipolar laser detector. Circuit can be used with CdSe cells or the PIN-10 board-area photodiode. With 3 type EM-1502 CdSe cells mounted in a parabolic head, laser pulses were detected at 820 feet.**

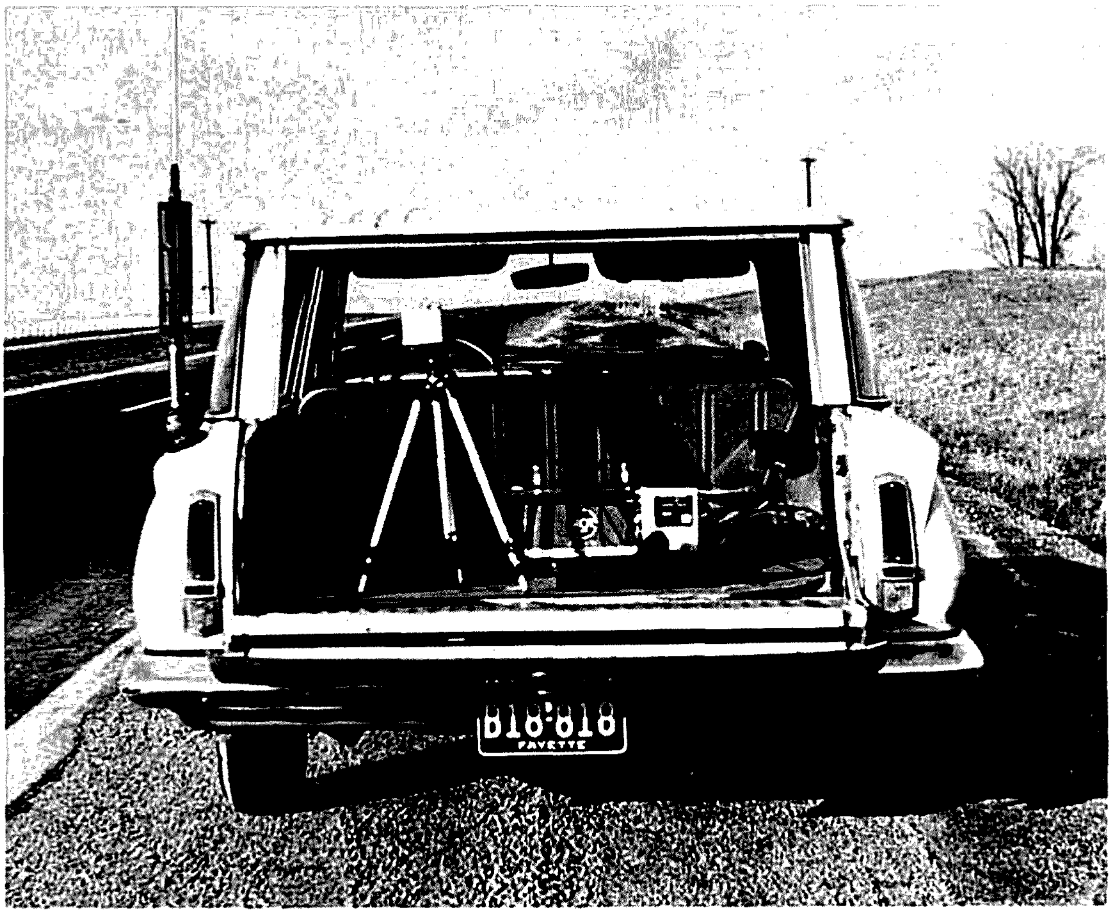
works. Regular-sized terminal strips are preferable for mounting the larger 0.01 and 0.005 $\mu$F capacitors. Epoxy was used to mount the scr in the pulser chassis.

## collimating optics

The M3 Sniperscope shown in the photo is used to collimate the near field of the laser and for boresighting laser and using the 1P25 and British CV-148 tubes aren't recommended for this use. If anyone knows of a surplus instrument using a 6929 or 6914 image tube, I'd appreciate receiving the information.

## range tests

A problem in testing this equipment is finding a clear, flat, unobstructed area.



**Author's mobile ranging equipment. Maximum distance of 1/5 mile was obtained on a new section of highway, but not without some problems.**

My first tests were conducted at the local Bluegrass Airport, where I obtained permission to use the runway. While ducking airliners, I was able to achieve 430 feet range. This short range was discouraging; however, I didn't realize at the time that the Sniperscope image tube couldn't detect peak power. Such are the consequences of adversity. But I pressed on.

My next ranging attempt was made at a closed section of new highway. I experienced all sorts of problems here—not with the equipment, but with the local constabulary, earlybirds looking for shortcuts, an unmarked panel truck that put its spotlight on my distant vehicle (resulting in a rapid signal return) and other inconveniences. Fortunately my tests had been completed, resulting in a range of 820 feet.

The final tests resulted in 1120 feet, using the PIN-10 broad-area detector. My next project, still in the research stage, is to operate a keyed-pulse temporary communications link over a carefully chosen route with another amateur using a pair of optical transceivers. Any volunteers?

references


1. R. W. Campbell, W4KAE, "Gallium Arsenide LED Experiments," *ham radio*, June, 1970, p. 6.
2. C. H. Gooch, "Gallium Arsenide Lasers," John Wiley & Sons, 22 Worcester Road, Rexdale, Ontario, Canada.
3. "Silicon Photodetector Calculator," United Detector Technology, 1732 21 St., Santa Monica, Calif. 90404.
4. Data sheet, "Gallium Arsenide Laser Diode Array LD200, LD200S Series," Laser Diode Laboratories, 205 Forrest St., Metuchin, N. J. 08840.
5. Data sheet, "RCA Gallium Arsenide Injection-Laser Diodes, TA2628, TA2628R," Solid-State Optical Engineering, RCA, U. S. Route 202, Sommerville, N. J. 08876.
6. "Pulsing the Vest-Pocket Laser," *Optical Systems Design*, July/August, 1969.


ham radio

# frequency spotter

## for general-coverage receivers

A marker generator featuring a sure-fire oscillator circuit and easily obtained parts

Carl C. Drumeller, W5JJ, 5824 N. W. 58th Street, Warr Acres, Oklahoma 73122

**Many amateurs have** a general-coverage receiver as well as an amateur-band-only receiver or transceiver. This is a desirable item, filling in those in-between band spots for casual listening or for checking for out-of-band radiation. A weak point for most such receivers is that they're usually of a less expensive type and thereby lack the precise frequency read-out you'd like to have.

This lack of precise frequency calibration complicates the use of the more common type of calibrator. A 100-kHz calibrator can cause more confusion than it can resolve when dial calibration on a receiver is such that you can't identify which of the marker signals represents the desired reference signal. In such instances a 1-MHz spotting signal is much more convenient to use. Most often, too, 1-MHz markers are sufficient for setting the bandset dial before using the calibrated bandspread dial.

There are many circuits for low-frequency crystals (100-200 kHz) and even more for the high-frequency spectrum. These circuits, unfortunately, don't always perform satisfactorily in the medium-frequency range.

This article describes a circuit especially adapted to a 1-MHz crystal. It oscillates dependably and produces profuse harmonics to the upper limit of the high-frequency band. Best of all, it can be built with junk-box parts with no requirement to purchase some hard-to-find component.

### construction

As the schematic shows (fig. 1), this circuit borrows liberally from the Miller, Colpitts, and Pierce. The coil in the collector-to-emitter circuit suggests the Miller; the use of a capacitive voltage

divider in the collector-emitter-ground string is strongly reminiscent of the Colpitts; the crystal is placed between the collector and base, as in the Pierce. Which takes ascendancy, I don't know. But it oscillates with no hesitancy, and that's the reason why I selected it.

At this frequency, the arrangement of components isn't critical. I used a small piece of perf board and mounted parts with no thought to short leads. The only item I mounted with care was the inductor, which I placed away from and at right angles A to the rf choke coil. The transistor was straight out of the junk box; before hitting the junk box, it lived on a printed-circuit board native to some of the receiver) there's ample room to mount a miniature toggle switch under the headphone jack. This switch must be mounted for a back-and-forth movement (not up and down) to have sufficient clearance. This coincides nicely with the movement of the slide switches used elsewhere.

Only the receiver's bottom plate needs to come off to permit mounting the switch and the oscillator board. There's plenty of room in the underside of the cabinet to mount the oscillator board. The chassis serves as the negative power lead, and the positive lead can be tapped off the bus at the main filter capacitor. As the set's voltage is a little higher than

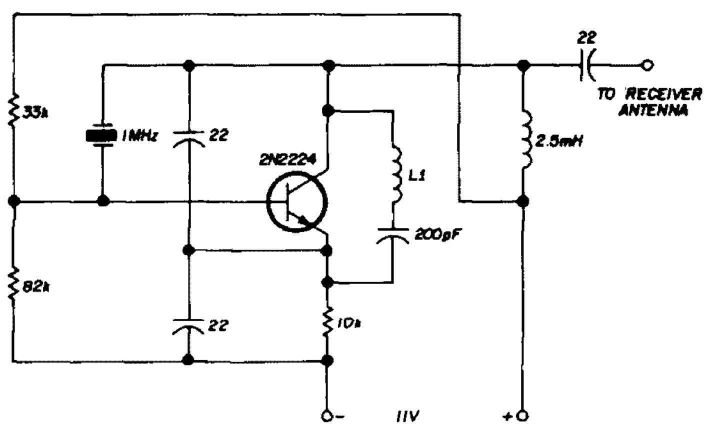



**fig. 1. Schematic of the frequency spotter. A 1-MHz crystal works best in this application. L2 is 80 turns no. 32 scramble wound on a ½" form.**

unidentified equipment. The resistors and capacitors also came from various printed-circuit boards. The rf choke, I regret to say, seldom appears in present-day surplus offerings; I had to buy mine, a distinctly unpleasant transaction! The coil was a junk-box progeny, too. The ½-inch form came from an old TV set, and the no. 32 wire was salvaged from an ancient electrodynamic speaker field coil. I doubt that the value of any component is critical; most likely, any nearly alike substitute would function equally well.

## installation notes

The receiver in which I placed my calibrator is a Radio Shack offering, the DX-150. This receiver lends itself well to the addition of a frequency spotter. On its left-hand side (looking from the front that demanded by the spotter, I used a 2000-ohm resistor in series with the positive power lead. This serves the dual purpose of dropping the voltage and insuring against zapping the receiver's power supply in case of a short circuit in the oscillator board.

Whatever receiver you use, you'll no doubt find a ready means of mounting your frequency spotter. If your receiver uses vacuum tubes, don't overlook the ready source of transistor power available from the cathode end of the cathode-bias resistor in the audio power amplifier stage. It's well filtered and usually just about the voltage you're looking for.

This is a pleasing little device. You'll enjoy building it, and you'll find it a much-used adjunct to your receiver.

ham radio

# radio teletype

## using

## ssb transceivers

The advantages of transceiver operation for RTTY can be realized by using audio-frequency-shift keying

D. H. Phillips, W6FOO, 215 Carlisle NE, Albuquerque, New Mexico 87106

Many of us who work the hf ham bands are ssb transceiver operators. The pleasure of transceiver operation may spoil you, since your time is not consumed by making tuning adjustments. However, all is not perfect. Although transceivers are great for phone operation, and acceptable for cw operation, they can be used to transmit RTTY only if unusually good engineering practices are employed.

The FCC does not permit tone-modulated am on the hf ham bands. Consequently, fsk transmission must be accomplished by some carrier-shift technique. In the old days, when we all had transmitters and receivers, RTTY transmission was simple and easy. A "shift pot" circuit could be used to vary the transmitter vfo frequency to encode RTTY mark and space signals into fsk. Phase continuity of the fsk waveform resulted because the fsk switching signal usually had long rise and fall times (because of RC time constants) compared to the period of the vfo frequency. Independent tuning of the receiver allowed the mark and space tone filters in the receiving converter to be tuned to convenient frequencies, within the receiver audio passband, as long as the two frequencies were separated by the proper deviation (850 Hz for wide shift and 170 Hz for narrow shift).

### ssb transceiver rtty

With a single-sideband transceiver, RTTY operation is not so simple. Any vfo modification can affect the frequency stability and degrade the performance of the transceiver on both RTTY and ssb. In addition, a more serious problem associated with transceive operation must be considered. That is, if the vfo frequency is shifted when transmitting, but not when receiving, then retuning may be required to receive an RTTY station that returns your call on *your* transmit frequency. This situation results in both stations "leapfrogging" up or down the band as each station retunes the other during receive intervals.

One rather undesirable method for eliminating leapfrogging is to operate the transceiver with carrier insertion in the transmit mode and in the upper-sideband receive mode. The vfo frequency can be shifted so that the transmitted carrier frequency is shifted up 1.65 kHz for space and up 2.5 kHz for mark. The space carrier will then be transmitted on a frequency 850 Hz lower than the mark carrier, in conformance with standard practice.

Under these conditions, if the vfo frequency is not shifted at all in the

receive mode, the received tones will be 1650 Hz (space) and 2500 Hz (mark) when the received station is zero beat with your transmit frequency. This method isn't very practical and not widely used because of the difficulty in adjusting the carrier shifts. The adjustments must be made when the transceiver is in the transmit mode (receive section disabled); therefore, proper adjustment requires a second receiver.

## afsk principles

A far better technique for transmitting RTTY with an ssb transceiver involves the use of audio frequency shift keying (afsk). A transceiver having good primary-carrier suppression can be operated in the ssb mode and modulated with an audio tone. If the primary carrier is adequately suppressed, a listener who tunes his receiver to such a signal hears only a pure carrier. This secondary carrier is displaced from the primary carrier by the audio frequency.

Using this principle, 850-Hz carrier-shift fsk can be transmitted with an ssb transceiver modulated by an audio tone shifted by 850 Hz to encode the mark and space information. If an ssb transceiver is operated in the usb mode and modulated with space and mark tones on 1650 and 2500 Hz, the transmitted space carrier would be 850 Hz below the mark carrier, in conformance with standard practice. In the receive mode, 1650- and 2500-Hz space and mark tones are received from a zero-beat RTTY station.

This popular method is used by many hams because it takes advantage of transceive operation by tracking transmitted space and mark tones. System alignment is straightforward, and no transceiver modifications are required. Although this method is very attractive, good design practices must be observed to ensure a high-quality fsk signal. Problems that must be considered for this method to be used successfully are discussed next.

## sideband and carrier suppression

The transmission of two carriers simultaneously is not permitted by the FCC. Consequently, the ssb exciter must suppress the unwanted sideband and the primary carrier by at least 30 dB, and the exciter must operate with the primary carrier completely balanced out. With 30-dB carrier suppression, a transceiver with 500 watts of secondary carrier (fsk signal) will be transmitting only one-half watt of primary carrier. This is a safe value and won't alarm adjacent channel users or the FCC.

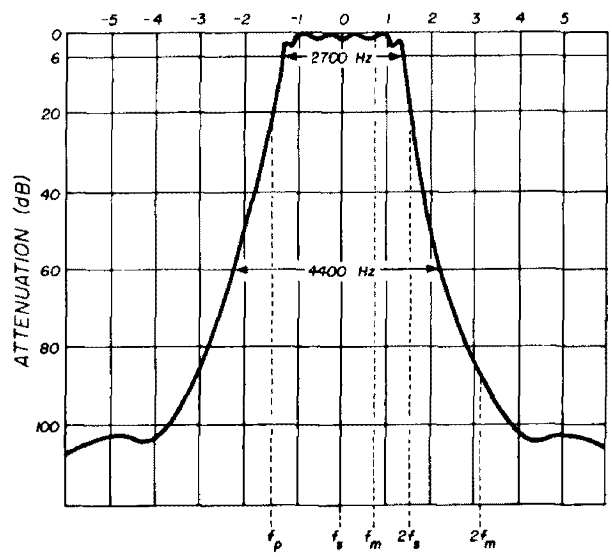



**fig. 1. Typical bandpass characteristic for a filter-type ssb (usb) exciter in the RTTY mode. Primary carrier is represented by fp; secondary carriers of 1650 Hz (space) and 2500 Hz (mark) are shown by fs and fm.**

## the audio signal

Any noise accompanying the audio tones will be transmitted, resulting in a waste of power and causing adjacent-channel interference, which is illegal. The signal-to-noise ratio of the audio-tone signal that modulates the transceiver should be at least 35 dB to ensure against transmitting appreciable noise.

The audio tone waveforms should be high-quality sinusoids having less than 5% distortion. A nonsinusoidal waveform contains harmonic components of the fundamental frequency. For example if a nonsinusoidal 500-Hz tone modulated the transceiver, several undersirable secondary carriers would be transmitted, which could have frequencies displaced from the

primary carrier by 500, 1000, 1500, 2000, and 2500 Hz.

The crystal- or mechanical-filter bandpass characteristics in filter-type ssb transceivers would attenuate any secondary carrier displaced from the primary carrier frequency by more than 2.7 kHz. Multiple secondary-carrier transmission is not allowed by the FCC, so a high-quality sinusoidal audio modulation waveform is required.

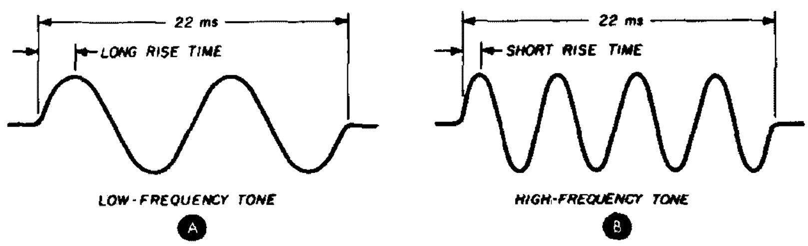



**fig. 2. Comparison of low- and high-frequency afsk tones. Signal at B allows use of more selective filters and provides a better leading edge of space or mark signal.**

## mark and space tones

The mark and space tone frequencies can be selected to compensate for deviations from a pure sinusoidal audio waveform. **Fig. 1** shows a typical crystal-filter *bandpass characteristic of an ssb trans*ceiver manufactured for ham use. Shown is the position of the primary carrier, $f_p$, relative to the filter passband. With the primary carrier on the skirt of the filter selectivity curve, the crystal filter attenuates the primary carrier and provides more carrier suppression than can be obtained in the balanced modulator alone. The tone frequencies are chosen so that the associated secondary carriers will be near the high-frequency cutoff point, but within the flat-response region.

**Fig. 1** shows the relative position of the primary carrier, $f_p$, and the secondary carriers, $f_s$ and $f_m$, for space and mark tones of 1650 and 2500 Hz. These fundamental space and mark frequencies lie within the passband and are not attenuated. However, the harmonics of the fundamental tones lie outside the passband and will be attenuated. Consequently, the high-frequency-filter selectivity skirt attenuates the secondary carriers produced by any existing harmonic content (distortion) of the audio tones.

**Fig. 1** also shows that by selecting 1650-Hz space and 2500-Hz mark frequencies, the second-harmonic space and mark secondary carriers, $2f_s$ and $2f_m$, are attenuated by 20 and 85 dB respectively. Although this technique can be used to compensate *partially for* audio tones having some harmonic content, every attempt should be made to generate pure sinusoidal modulation-tone waveforms.

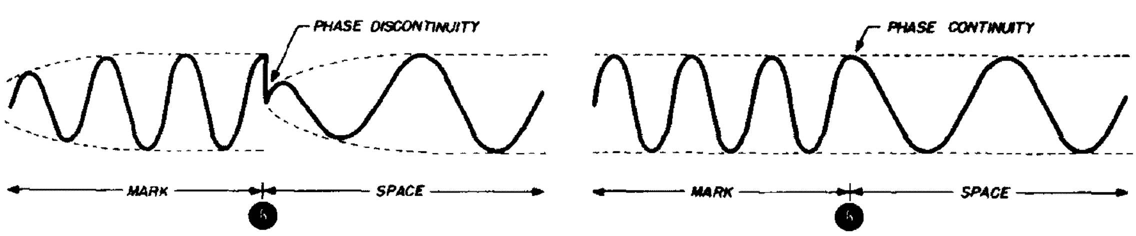



**fig. 3. Comparison of mark-to-space wave-form transitions. Phase continuity avoids adjacent-channel interference because impulse noise is greatly reduced.**

There is a second reason for choosing tone frequencies near the high-frequency skirt of the filter-selectivity curve. High tone frequencies are desirable so that more cycles of the tone can be contained within the 22-millisecond space and mark intervals. High tone frequencies allow the

use of more selective tone filters in the receiving converter. Also, a high tone frequency represents a shorter rise time, or a better leading edge, of the space or mark signal, as shown in **fig. 2**.

## phase continuity

Finally, perhaps the most important consideration is the requirement for generating a continuous-phase afsk signal. The need for phase continuity is best emphasized by considering the effects of an afsk modulation waveform having phase discontinuities. **Fig. 3** illustrates the difference between the two waveforms. The waveform with phase discontinuity results in (a) amplitude modulation of the fsk signal, and (b) impulse noise being generated and transmitted.

The sharp switching transients have many frequency components, resulting in the transmission of a 3-kHz-wide noise spectrum. The transmitted noise bandwidth is limited by the crystal or mechanical filter frequency response. Because the duty cycle of the switching transient is low, the transmitted noise power is much less than the secondary carrier power. However, this impulse noise does cause adjacent channel interference.

## measurements

Comparison measurements were made to evaluate continuous- and discontinuous-phase afsk modulation waveforms. The experimental setup for measuring the transmitted power spectrum is shown in **fig. 4**.

**Fig. 5** shows an oscillogram of a discontinuous-phase afsk waveform having a switching frequency of 100 Hz. This switching frequency, which is somewhat faster than the 60-wpm teletype switching frequency of 22.8 Hz, was used for convenience. The waveform of **fig. 5** was used to modulate a Swan 500 ssb transceiver. The Swan 500 was intentionally adjusted for a 3600-kHz primary carrier suppression of only 30 dB, so that the low-power primary carrier could be used as a reference signal on the spectrum analyzer trace. **Fig. 6** is an oscillogram of the spectrum analyzer display. (The blip on the leading edge at the extreme left of the trace was generated within the measuring instrument and doesn't indicate a lack of lower-sideband suppression.) **Fig. 6** shows that the phase discontinuities in the modulation waveform contribute to unequal space and mark power distributions. Impulse noise, although present, can't be seen on a filter-envelope spectrum analyzer trace.

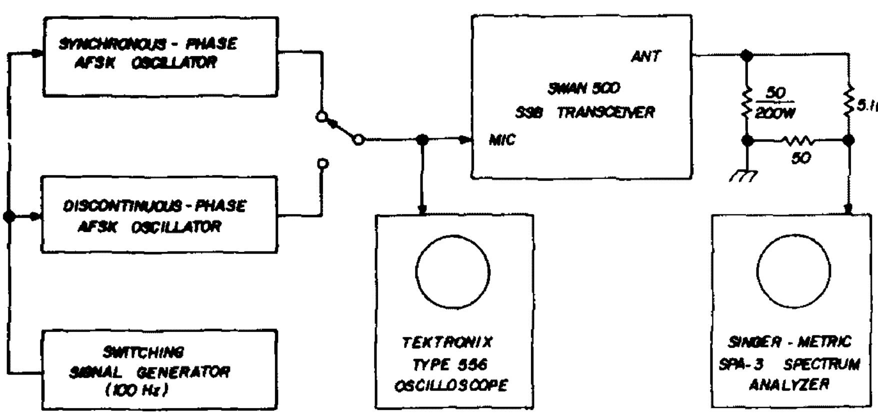
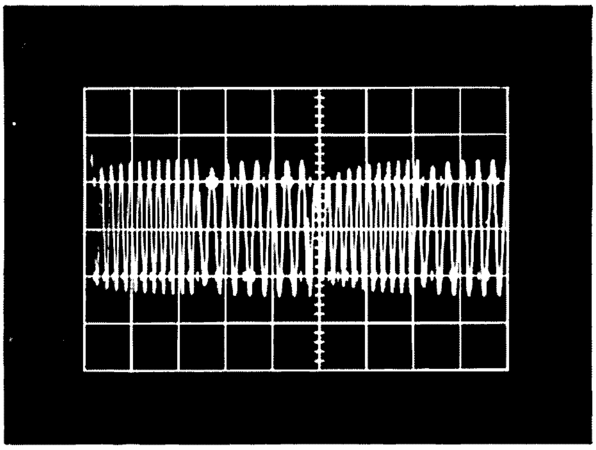



**fig. 4. Setup for measuring output-power spectrum of a transceiver modulated by a synchronous- and discontinuous-phase afsk oscillator.**

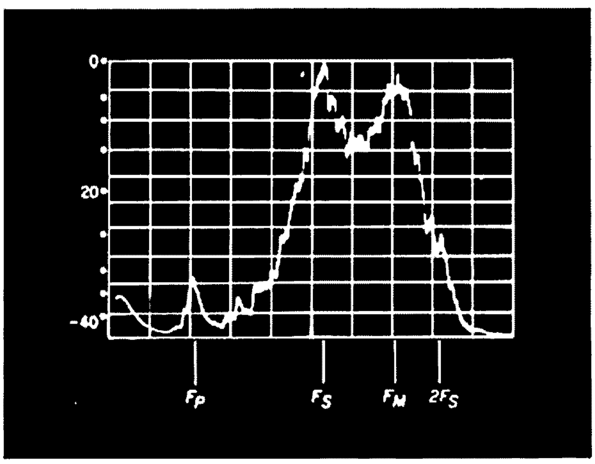
**Fig. 7** is an oscillogram of a synchronous-phase afsk waveform, which was used to modulate the Swan 500. The resulting power spectrum, shown in **fig. 8**, illustrates the space and mark power distributions and the low harmonic generation that can be obtained by using a synchronous-phase afsk oscillator. A comparison of **figs. 5** and **6** with **figs. 7** and **8**



| | |
|---|---|
| Space Tone | 1650 Hz |
| Mark Tone | 2500 Hz |
| Switching Freq | 100 Hz |
| Vertical | dB |

**fig. 5. Afsk waveform containing phase discontinuites at mark-to-space and space-to-mark transitions.**

emphasizes the improvement in RTTY signal quality that can be obtained by using synchronous-phase afsk modulation.

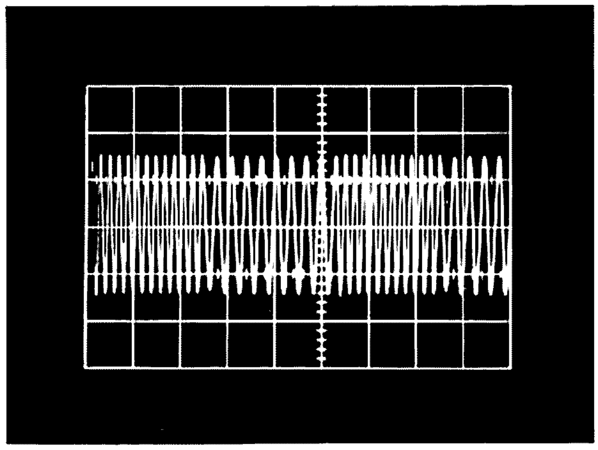



| | |
|---|---|
| Space Tone | 1650 Hz |
| Mark Tone | 2500 Hz |
| Switching Freq | 100 Hz |

**fig. 6. Spectrum-analyzer display of power spectrum from an usb exciter modulated by a discontinuous-phase afsk signal.**

## summary

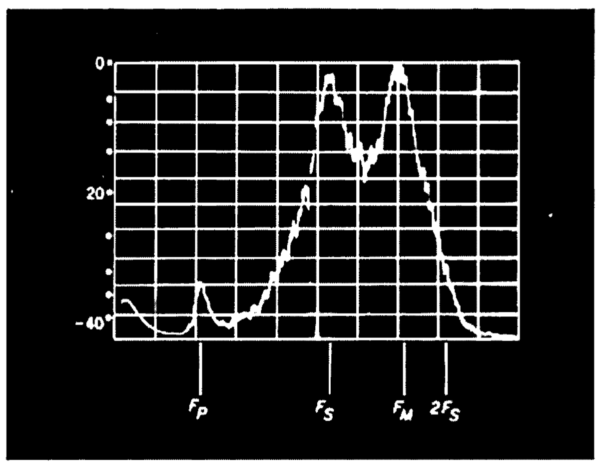
Properly selected audio-tone frequencies and synchronous-phase afsk modulation can be used to generate high-quality pseudo-carrier-shift RTTY signals with ssb exciters or transceivers. Perfect-



| | |
|---|---|
| Horizontal | kHz |
| fp = 3600 kHz | |
| fs = fp + 1650 Hz | |
| fm = fp + 2500 Hz | |

**fig. 7. Synchronous-phase afsk waveform.**

phase-continuity modulation waveforms will minimize adjacent-channel interference and result in a better contact record by improving the printer copy at




| | |
|---|---|
| Vertical | 10mV/div |
| Horizontal | 2ms/div |

**fig. 8. Power spectrum from an usb exciter modulated by a synchronous-phase afsk signal.**

the other end of the radio circuit. These improved RTTY transmission techniques will also eliminate FCC citations.

A practical synchronous-phase afsk oscillator using an inexpensive IC flip-flop will be described in a future article.

ham radio

# auxiliary receiver for 160 meters

Easy and inexpensive conversion of an auto radio for emergency or standby service

F. J. Bauer, Jr., W6FPO, P. O. Box 870, Felton, California 95018

If you have an idle broadcast-band auto radio lying around, why not put it to work as an emergency or standby receiver for 160 meters? It can be converted quickly with a minimum outlay for parts. For a-m reception, all you need are an oscillator coil and a couple of inexpensive adjustable rf chokes. With a few more components you can add circuits for cw and ssb.

## basic conversion circuit

The basic conversion consists of reducing the front-end coil inductances with shunt coils so the receiver will cover 1.8 to 2 MHz (fig. 1). If the alignment instructions are followed, this conversion method will result in much better performance than if the receiver were merely misaligned to cover part of the band.

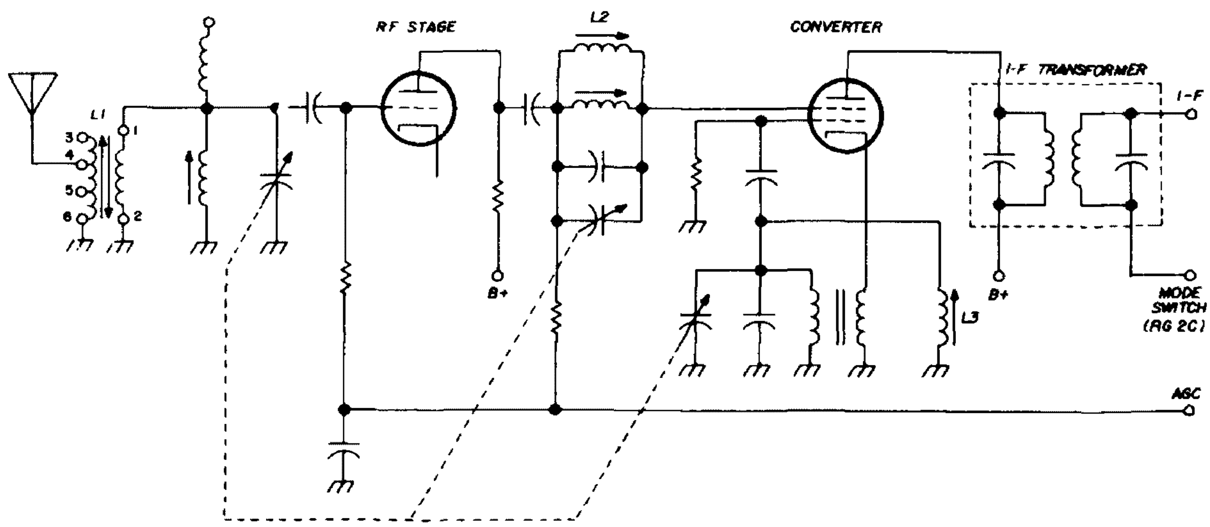
To prepare the receiver for conversion, first check its performance "as is." The receiver can be tested on the bench using an ac power supply after removing the vibrator. Check alignment and make any required repairs. When you're satisfied the receiver is performing as it should, replace the vibrator transformer with a 60-Hz unit of the proper rating.

Next install the rf coils and connect them directly to the tuner coil terminals. The auxiliary coils are designed for single-hole mounting and can be installed anywhere as long as you have access to the slugs. The only precaution is to isolate the antenna coil from the rf stage, or the latter will oscillate. I solved this problem by mounting the antenna coil on the opposite side of the chassis from the rf and oscillator coils.

sponse (dial should be set to about 900 kHz).

6. Repeat steps 3, 4, and 5.

The oscillator-coil slug is adjusted until you hear the oscillator signal in another receiver or frequency meter. Lacking this equipment, the oscillator frequency can be adjusted by injecting an 1800-kHz signal into the converter grid. This is easily done by tuning a bc transistor radio




| | | | |
|---|---|---|---|
| L1 | Adjustable BC band oscillator coil, approx. 190-360 μH (J. W. Miller 71-OSC) | L2 | 50-140 μH (J. W. Miller 4207) |
| | | L3 | 10-25 μH (J. W. Miller 4205) |

**fig. 1. Basic conversion for 160-meter a-m reception on a bc-band auto radio. Front-end tuned circuits are shunted with outboard coils to lower inductance. Miller 4205 coil can be adjusted for either 455- or 262-kHz i-f.**

## alignment

The easiest way to align the receiver is as follows:

1. Set receiver dial to 600 kHz.
2. Adjust oscillator coil to:
   a. 2062 kHz (receiver i-f = 262 kHz).
   b. 2255 kHz (receiver i-f = 455 kHz).
3. Adjust rf and antenna coil slugs for maximum response to an 1800-kHz signal.
4. Set signal source to 2000 kHz.
5. Tune in signal on receiver and adjust trimmers for maximum re-

to 1340 or 1350 kHz and picking up the oscillator radiation as an 1800-kHz signal. For an approximate 2000-kHz signal source (step 4 above), you can tune the transistor radio to 1540 or 1550 kHz and pick up its mixer signal. Final retouching adjustment of the antenna trimmer with the antenna connected should provide good tracking over the band.

## cw and ssb reception

It will be necessary to add a bfo and manual rf gain control for cw and ssb reception (fig. 2). The bfo is powered by the output-tube cathode resistor as shown. A convenient way to mount the gain control and bfo switch is on the speaker cover of the set. All leads to these

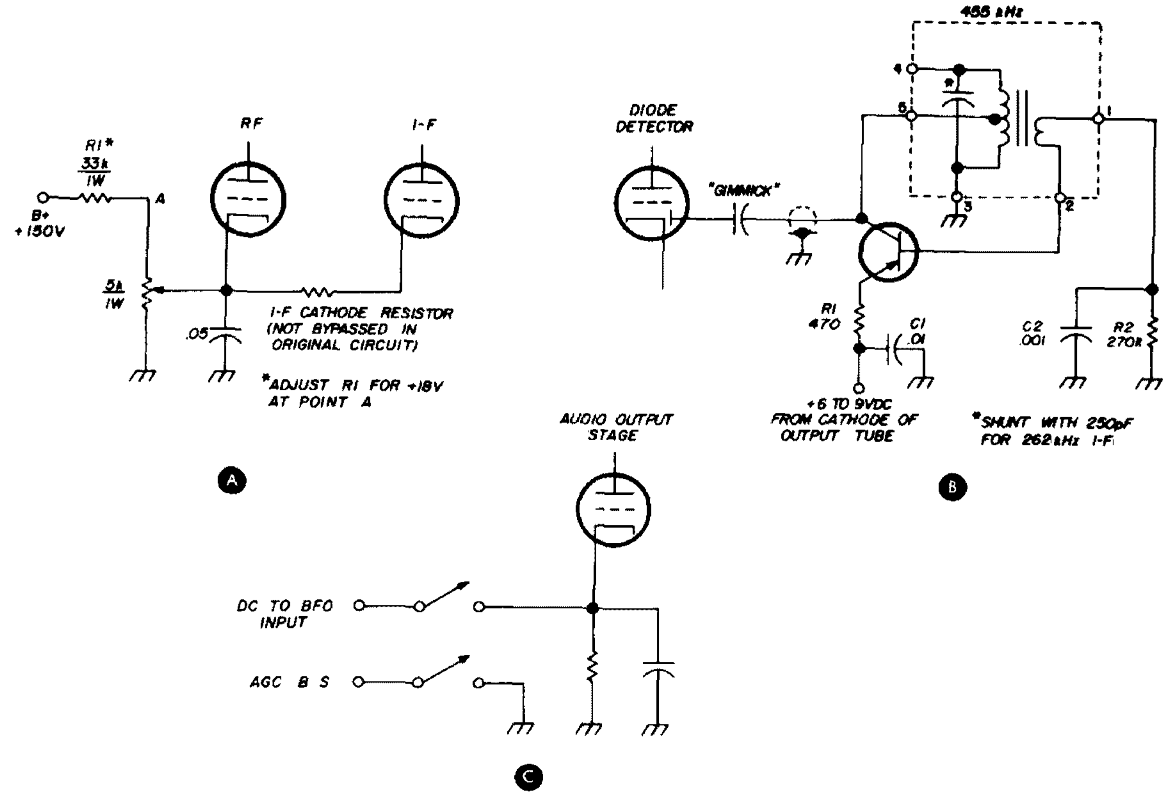



fig. 2. Additional circuits for cw and ssb. Manual rf gain control is shown in A; B and C show bfo and mode switch. The 455 kHz i-f transformer is a miniature transistor type such as the J. W. Miller 2042. Transistor in B is a small-signal pnp type.

circuits carry only dc, so make them long enough so the receiver can be easily serviced.

Note that the bfo switch also disables the receiver agc when the bfo is on. For cw or ssb reception back off the rf gain, set audio gain near maximum, and turn on the bfo switch. This may not be the most elegant way to receive ssb; but considering its simplicity and low cost, it's adequate for this type of receiver.

Almost any small-signal pnp transistor will oscillate in the bfo circuit. For my conversion, the i-f transformer and transistor were lifted from a defunct transistor radio. If the thing won't oscillate, try reversing the i-f transformer secondary leads. If this doesn't work, try another transistor.

## bfo adjustment

Before permanently wiring the bfo to the switch, turn on the bfo with receiver agc activated. Adjust the bfo tuning slug for maximum agc voltage. This centers the bfo in the i-f passband. No bfo adjustment control is provided, because the i-f in these old receivers is so broad that final pitch adjustment can be made with the tuning dial.

For proper injection, the bfo output is coupled to the diode detector through a shielded gimmick. The coupling should be adjusted so that the agc voltage increases by about 0.3 volt when the bfo is turned on. Excessive coupling will overload the detector and may cause motorboating when the audio gain is turned all the way up.

The bfo components can be mounted on a 1- by 2-inch perforated board. The transistor, R1, and C1 are mounted on the left side of the transformer, and C2, and R2 are mounted on the right side. The output lead shield should be grounded at the bfo board. This will ensure against bfo radiation being picked up by the i-f stage, which causes noise.

This conversion admittedly uses none of the latest sophisticated techniques for superb cw and ssb reception, but it does provide a good emergency or standby receiver for minimum effort and cost.

**ham radio**

# a counter gating source

The power-line frequency makes an accurate time base for many counter applications

E. H. Conklin, K6KA, Box 1, La Canada, California 91011

**Several articles on** simple frequency counters have suggested using the 60-Hz line frequency as a gating source. One commercially built unit uses it, claiming 0.02-percent accuracy compared with the ±1-count accuracy of a temperature-controlled crystal source.

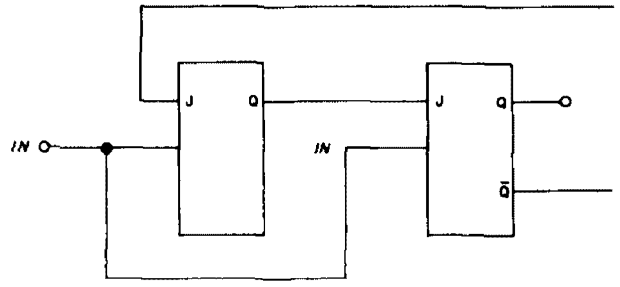
The ± 1-count accuracy is desirable for the ARRL frequency-measuring tests, or when a calibrator is desired for 10, 5, or even 2.5-kHz markers. For other applications, such as a digital counter dial for a transmitter or receiver, the error resulting from using the 60-Hz line frequency shouldn't appreciably affect the final tenth-kHz digit in the readout. Therefore, the idea of the resulting simplification in equipment is intriguing.

Jim Fisk once pointed out that the power-line frequency may follow different patterns in different locations. However, power networks in most parts of the country are quite large, and sophisticated frequency regulators are used. Electric clocks, for example, are quite accurate when driven by power sources in most cities. With this in mind, I decided to get a little first-hand experience with the problem.

## 60-Hz time base

It is possible to count line frequency in several ways with considerable accuracy. I decided to use the 60-Hz line frequency as a time base to count the

**fig. 1. Test circuit using a Fairchild 9093 JK flip-flop as a divide-by-three counter to measure 60-Hz line frequency.**



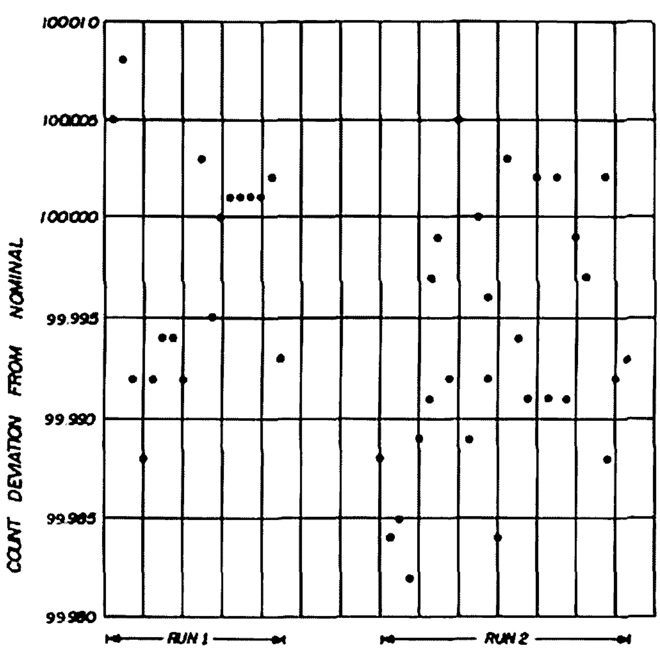
100-kHz output from a temperature-controlled frequency standard.[1,2] The result would be the accuracy of the power-supply time base expressed as a percentage carried out to three decimal places.

## test equipment

I fed 0.5 volt at 60 Hz into a Fairchild 9093 JK flip-flop connected in a divide-by-three circuit (fig. 1). No Schmitt trigger or other wave-shaping circuit was used. The counter correctly measured the 60-Hz line frequency about 80 percent of the time. During the other 20 percent of the time, one extra count appeared in the output. This is better performance than that specified for most counters using a Schmitt trigger or other wave shaper. The variation affects the units decade only.

The counter's 20-kHz output was then fed into part of the time-base board using only a divide-by-two flip-flop and one decade, producing 1-Hz, which operated the gate.[2] The 100-kHz calibrator crystal output was then counted. The result was in five or six digits, which can be taken as a percentage of the 60-Hz line frequency.

fig. 2. Scatter plot of results from two consecutive tests using 60-Hz line frequency as a gating source for a precision counter. Measurements were taken about two seconds apart.




## measurements

The results of two measurement runs are plotted in fig. 2. Run 1 contains the only cases of successive identical data (the four points at 100.001). Considering all the data, the final digit may be correct, or it may be one count high. This performance is inferior for ARRL frequency-measuring tests up to 14 MHz where the maximum error is ±1 Hz with the crystal generating the time base;

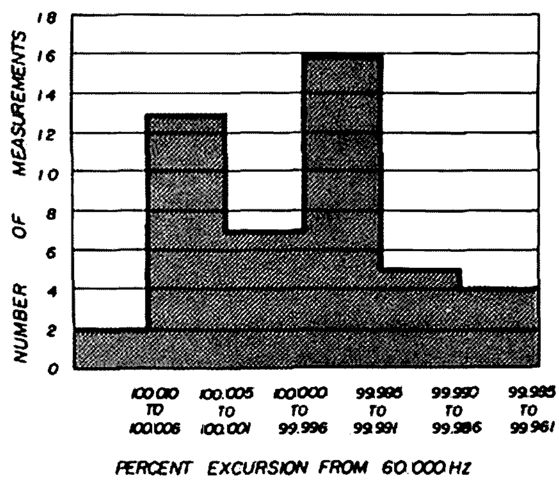



fig. 3. Histogram of measurement deviations from nominal 60-Hz line frequency.

however, the results were better than the 0.02-percent accuracy expected.

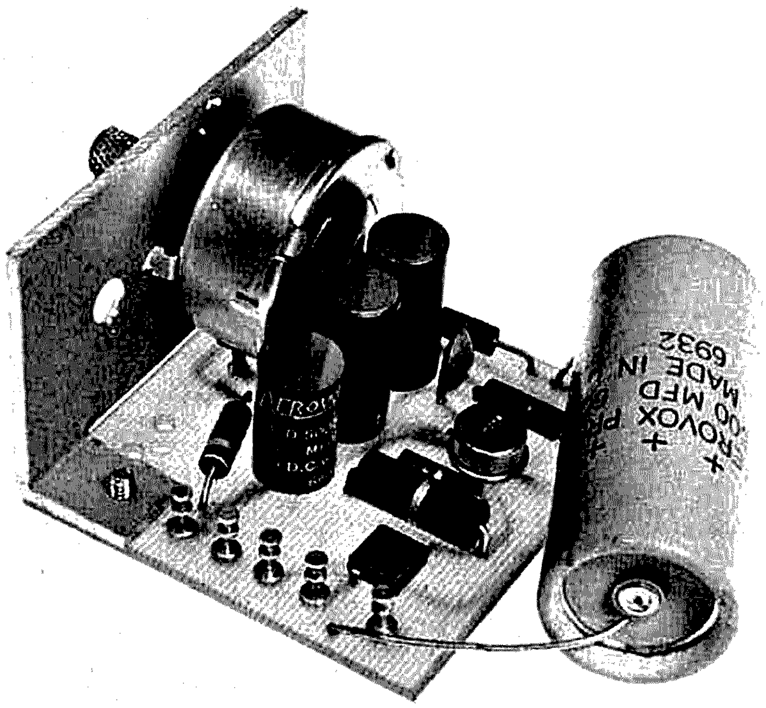
Examination of the data shows a count ranging from 0.018 percent low to 0.010 percent high (corresponding to points at 99.982 and 100.010, respectively, in fig. 2). The exact number may be anywhere within this range, since the data are essentially random.

The number of measurements falling within 0.005 percent from 60.000 Hz is shown in fig. 3. This suggests a bimodal distribution about the central value of 60.000 Hz.

## references

1. E. H. Conklin, K6KA, "Amateur Frequency Measurements," *ham radio*, October, 1968, pp. 53-54.
2. E. H. Conklin, K6KA, "Calibrators and Counters," *ham radio*, November, 1968, pp. 41-54.

ham radio




# voltage regulation— circa 1970

## A new unitized power supply using the CA-3055 integrated circuit

E. L. Klein, W2FBW

One of the most useful solid-state devices appearing on today's market is the integrated circuit designed for voltage-regulator service. Whereas a score of discrete components formerly were required to obtain a fraction of a volt of regulation, one small device now provides a ten-fold improvement.

A good regulated power supply is essential for experimenting with solid-state circuits. If you have yet to experience the advantages of solid-state technology, this article is for you. The circuits are extremely simple to build, and the experience gained will be helpful if you want to explore further in this area. Recommended reading on solid-state power supplies is an article by Hank Olson, W6GXN.[1]

## the IC regulator

The heart of this modern unitized power supply is RCA's new monolithic approach for two practical power supplies: one that will regulate load currents up to 100 mA, and a second supply that will handle even higher load currents.

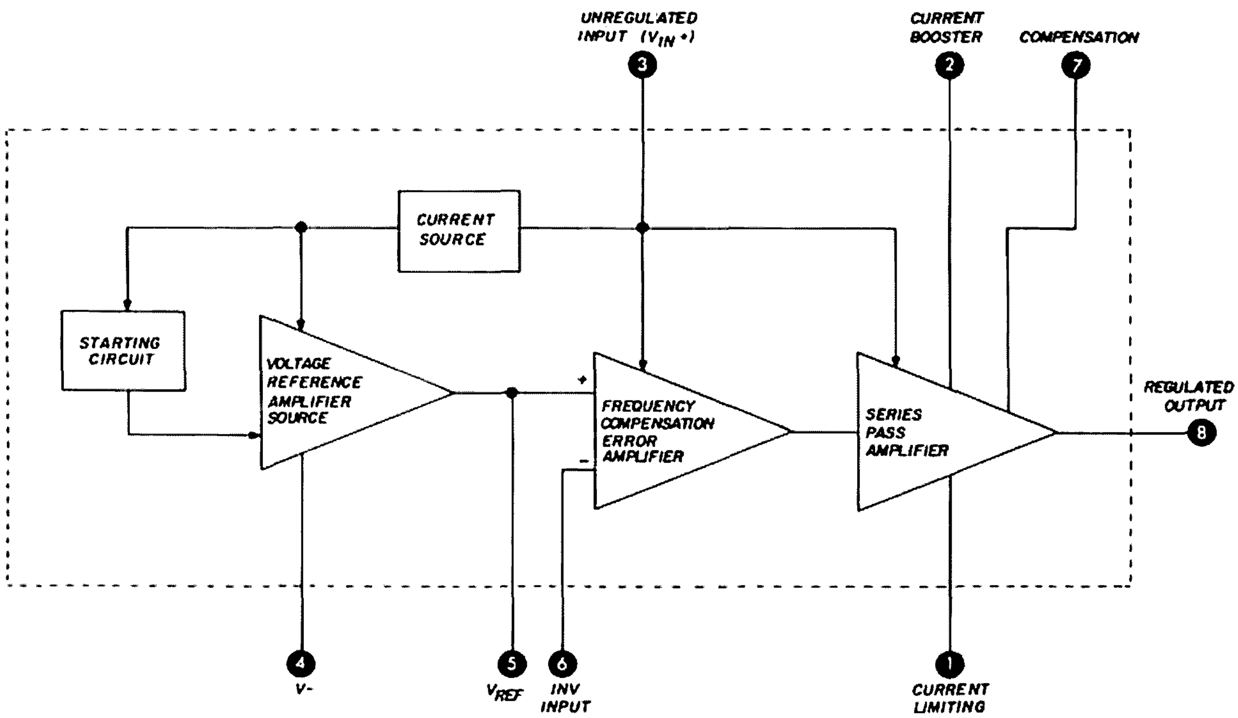



fig. 1. Block diagram of circuit elements in the RCA CA-3055 voltage regulator IC.

IC, the CA-3055 voltage regulator. Packaged in a TO-5 transistor case, the CA-3055 is less than a half-inch in diameter. It has eight leads that can be soldered directly into the circuit or to the terminals of a socket, such as the Cinch-Jones 8-ICS.

The regulator functions are shown in fig. 1. Also given are terminal designations for temperature-compensated reference voltage, booster input, frequency compensation, and short-circuit protection. The super-small construction allows over two-dozen components to be contained on a single silicon chip. Included are fifteen transistors, seven diodes, and four resistors.[2]

## unitized approach

An advanced design such as the CA-3055 deserves some ingenuity in its application. Therefore, I've included a new Here are the principal features of the unitized approach in regulated power-supply design:

CA-3055 integrated circuit.

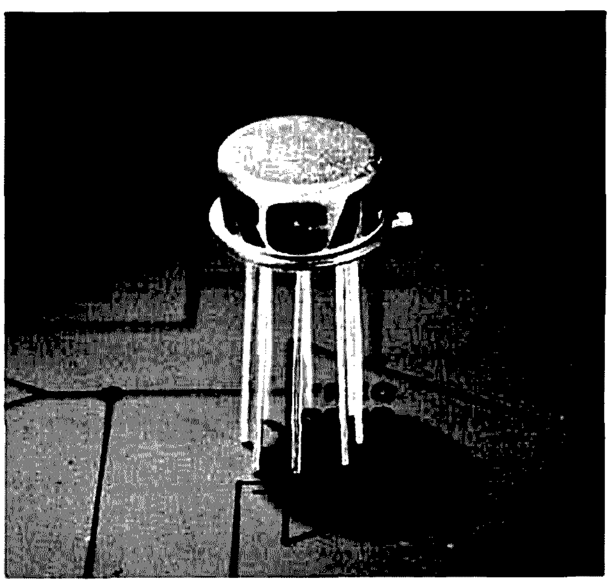


table 1. Output voltages using the CA3055, with input ac voltages supplied by various filament transformers.

| input voltage (Vac) | regulated output (Vdc) minimum | maximum |
|---|---|---|
| 6.3 | 1.75 | 4.25 |
| 12.6 | 2.15 | 10.0 |
| 25.2 | 2.50 | 18.0 |

1. **Single mounting hole.** This makes it easy to mount the power supply on the apron of an existing chassis, for example, or on the front panel of new equipment.

2. **No chassis.** Sheet-metal work is reduced to a minimum. All you need is a 3/32-inch-thick piece of aluminum, which is easily formed into a mounting bracket that serves as a circuit-board support and heat sink for the external pass transistor used in the high-current supply.

3. **Voltage control.** A small potentiometer may be included for proportional voltage adjustment—a desired refinement for many transistor circuits.

4. **Fixed voltage.** By selecting two fixed resistors in place of the potentiometer (and fixed resistors), discrete voltages are available. These may be switched or otherwise programmed for specific applications.

5. **Bench supply.** Add a voltmeter, milliameter, and an on-off switch, and a first-class bench power supply is ready to go.

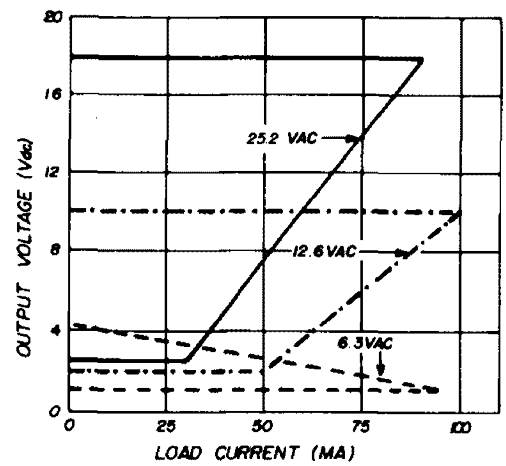
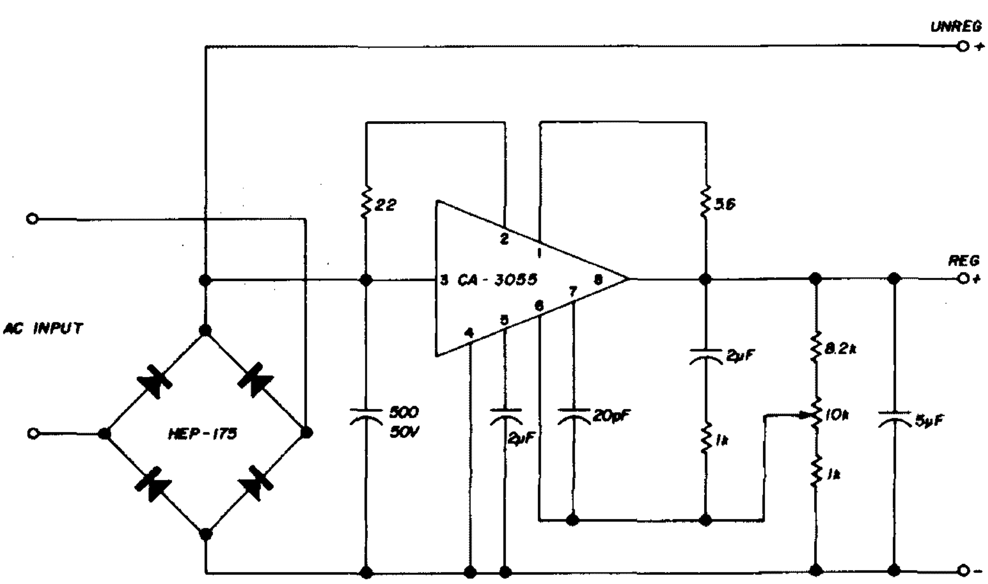


fig. 3. Useful ranges of the 100-mA unitized supply for three input voltages.

## practical supply for 100 mA

The CA-3055 can deliver output currents up to 100 mA without the use of external pass transistors. The IC's internal

fig. 2. Schematic of the 100-mA unitized power supply. Output ranges for three standard ac inputs are given in table 1.



series-pass amplifier handles this current. The only components needed other than the IC are the power source, rectifiers, filter capacitors, and voltage-adjusting resistors.

A schematic of the 100-mA supply is shown in fig. 2. Any standard filament transformer may be used as a power source.* The transformer, of course, will depend on your output-voltage requirements. Typical operating ranges of regulated dc output voltages for various filament-transformer secondary voltages are given in table 1. A plot of the supply's regulation characteristics for various input voltages is shown in fig. 3.

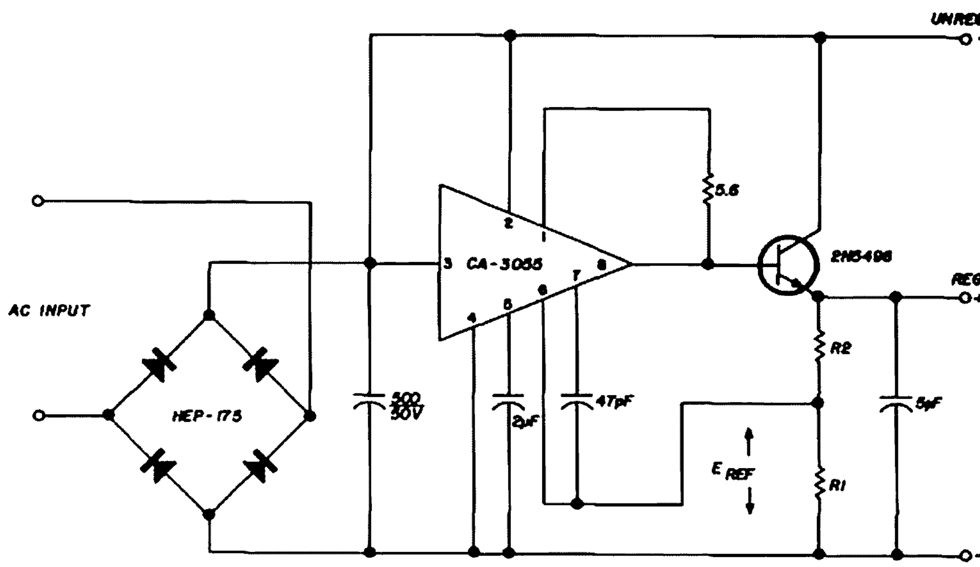
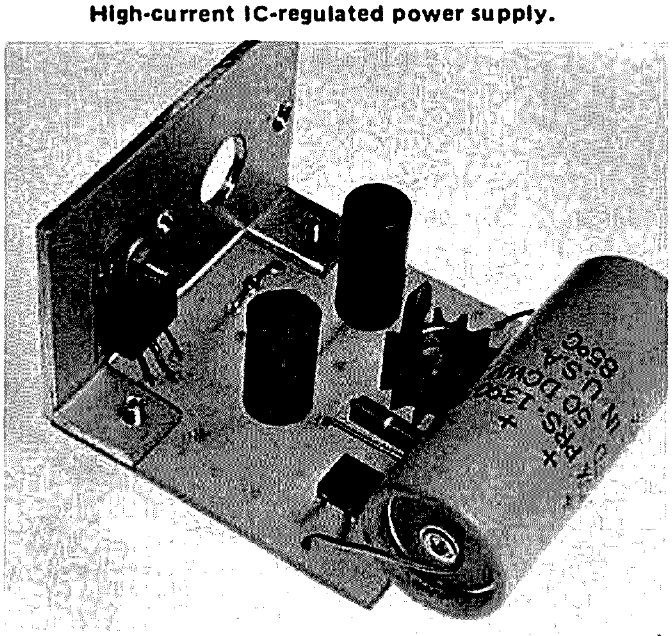



fig. 4. High-current supply. Load currents up to several amperes are possible.

High-current IC-regulated power supply.



## a high-current supply

The CA-3055 can also be used with a suitable external series-pass transistor to provide voltage regulation at loads greater than 100 mA. A typical circuit is shown in fig. 4. The 2N5496 transistor may be used with load currents up to several amperes in this circuit, provided the transistor is heat-sinked to the panel, as shown in the photo. Resistors R1 and R2 are selected for the required output voltage (table 2). If a continuously variable output voltage is desired the potentiometer control, as shown in the 100-mA supply, can be used.

## construction

All discrete components, as well as the IC and pass transistor, are mounted on a

*A Stancor type TP3 is suggested in RCA's data sheet for a typical supply (see reference 2). editor.

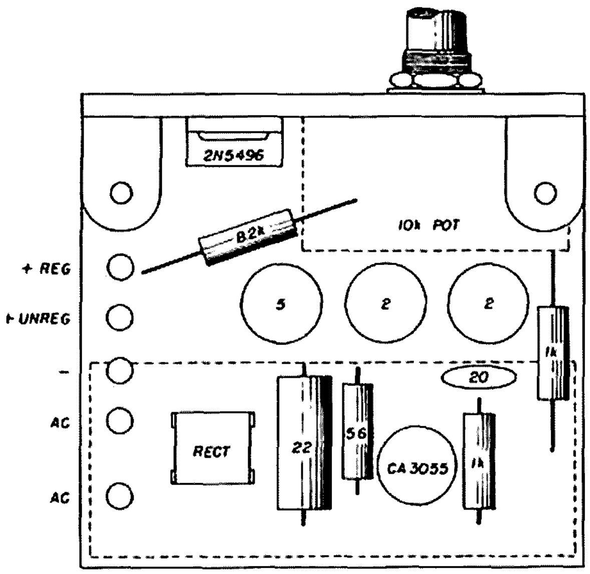



fig. 5. Layout of components on a circuit board.

table 2. Selections of resistors R1 and R2 for required output voltage in the high-current regulated supply.

| reg. output (Vdc) | operating range (12.6 Vac input) E ref (Vdc) | R1 | R2 |
|---|---|---|---|
| 15 | 1.5 | 1.7 | 17 |
| 12 | 1.5 | 2.1 | 16.6 |
| 10 | 1.5 | 2.5 | 16.2 |
| 8 | 1.5 | 3.2 | 15.5 |
| 6 | 1.5 | 4.3 | 14.4 |
| 4 | 1.5 | 6.4 | 12.3 |

$$\text{Reg (out)} = \text{E ref}\left(\frac{R1 + R2}{R1}\right)$$

2 x 2-1/8-inch phenolic board, fig. 5 and 6. The circuit board can be a prepunched board of the Vector board type, or can be built as a printed-circuit board. After all components are mounted and connected, the board is attached to the aluminum bracket (fig. 7) with small machine screws.

To enhance the appearance of the finished unit, I'd suggest sanding the bracket with medium-grit emery cloth, taking care to keep the scratch pattern parallel.

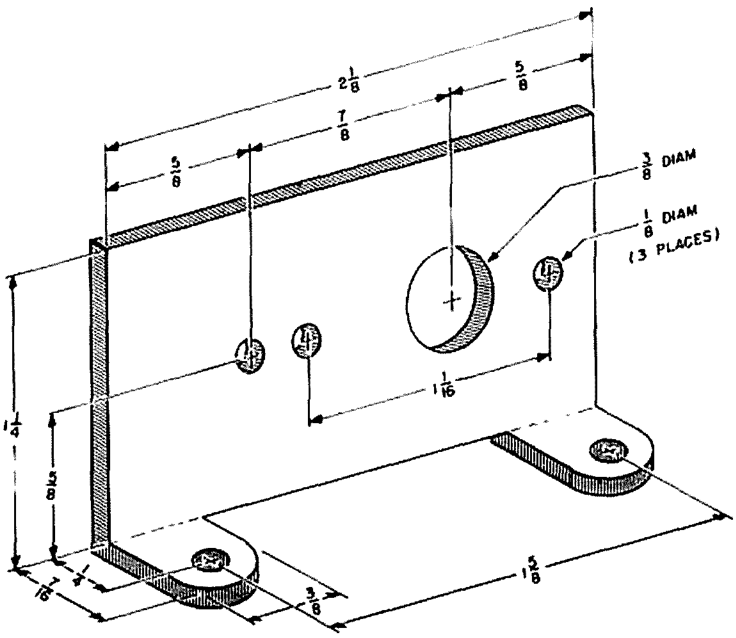



fig. 7. Aluminum mounting bracket.

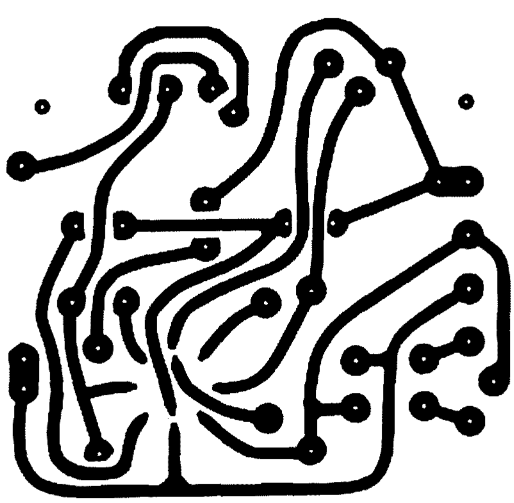


fig. 6. Printed-circuit board layout for the basic regulated supply.

Four different configurations for the unitized power supply are possible. Select the one to suit your needs. The few hours spent building this supply will be rewarding, and you'll be abreast of the times with your IC voltage regulator—circa 1970.

## references

1. Hank Olson, W6GXN, "A Survey of Solid-State Power Supplies," *ham radio,* February, 1970, p. 25.

2. "Linear Integrated Circuits CA3055," RCA data sheet file no. 395, RCA Electronic Components, Harrison, New Jersey 07029.

**ham radio**

# linear vhf tank circuits

## Parameters are presented for designing single-ended vhf tanks using nominal quarter-wave transmission lines

Harry Ferguson, K7UNL, 20629 5th Avenue South, Seattle, Washington 98148

**This article was prompted** by a search of various amateur publications and other literature for down-to-earth information on the design and construction of vhf linear tank circuits. I was interested in information that could be used by the amateur to design circuits for his particular needs rather than the "Chinese copy" type of article.

It all started when I decided that my 2-meter transmitter should be rebuilt to incorporate a single-ended linear tank circuit. Since practically all available design information on linear tank circuits was for push-pull output tubes, much time was spent on data research. I felt that the results of this effort, which are presented here, would interest the 2-meter enthusiast who wants to know a little more about the subject.

### equivalent circuits

Most single-ended vhf finals use the conventional parallel- or series-tuned inductance and capacitance plate circuit. The tank circuit discussed here is an adaptation of the parallel-tuned tank. It's basically a shortened ¼-wave coax line, short-circuited (for rf) at one end and resonated with a capacitance at the other.

This combination is equivalent to a parallel-tuned tank (fig. 1). Conventional single-ended circuits are also shown for comparison. The basic theory presented here can be applied to other types of linear circuits for higher frequencies, where the electrical length of the line is an integral number of quarter waves and the output-tube plate capacitance replaces part of one of the quarter waves. This discussion, however, is confined to ¼-wave lines only. Multiple ¼-wave lines used at frequencies higher than 145 MHz are discussed in detail in reference 1.

A coax line is unbalanced because of the way it's constructed. The outer conductor is at ground potential, and the center conductor is contained entirely within the outer conductor. Thus the electric field is confined completely within the shielded enclosure. This is ideal for a vhf tank circuit, since the tank provides its own shield, and radiation can be predicted and controlled. The design

problem is to choose a piece of transmission line that looks like an inductance and which forms a resonant circuit when connected in parallel with a capacitance.

the line's characteristic impedance and the tangent of the line's electrical length, $\phi$, in degrees, where 90 degrees = ¼ wavelength. The reactance is

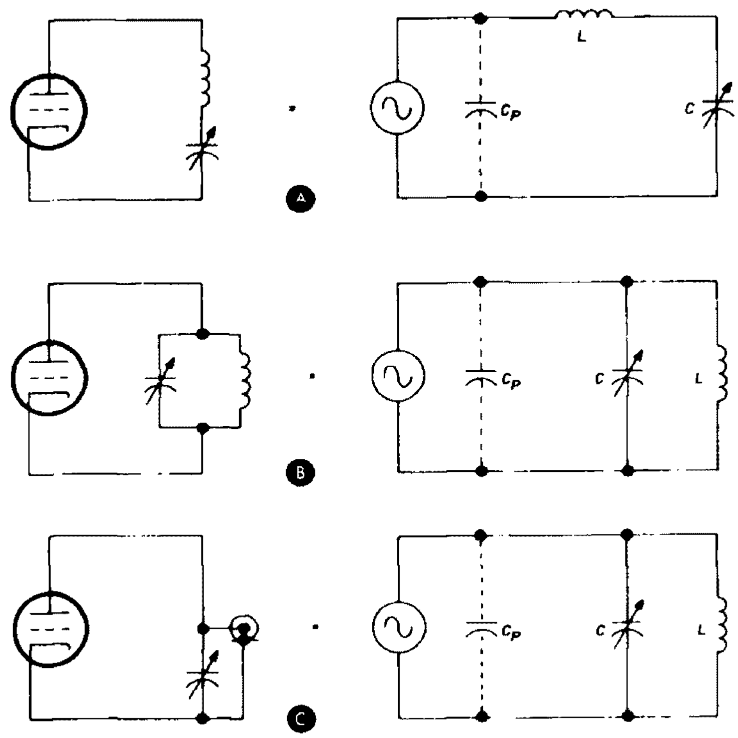



**fig. 1. Equivalent tank circuits. Conventional series- and parallel-tuned tanks using lumped constants are shown in A and B; linear parallel-tuned tank, C, is an adaptation of B.**

## background theory

At this point a property of transmission lines known as characteristic impedance must be introduced, since other factors in the design depend on the magnitude of this property. The characteristic impedance of a transmission line is a function of its physical dimensions (discussed in reference 2). I looked at four different transmission-line configurations to determine which would be easiest to build, consistent with the desired electrical characteristics. These are shown in fig. 2, which gives characteristic impedances for a range of practical values of conductor size and shape.

A coaxial line cut to a ¼ wavelength at a given frequency and short-circuited at one end presents an open circuit at the other end to rf at that frequency. Furthermore, for lengths shorter than ¼ wavelength, the impedance at the open end of the line is inductive. The magnitude of this impedance is the product of

$$X_L = Z_o (\tan \phi) \tag{1}$$

where

$X_L$ = inductive reactance (ohms)
$Z_o$ = characteristic impedance of line (ohms)
$\phi$ = electrical length of line (deg)

For shortened ¼-wavelength lines, the electrical length is always less than 90 degrees.

One step further into theory, and we find that electrical length is related to physical length by the constant 2952 for air-dielectric lines. This constant is equal to the velocity of propagation in air (in./µsec) divided by 4. So the length in inches of a ¼-wave line is obtained by dividing 2952 by the frequency in MHz. For any dielectric other than air, the length is reduced by the propagation constant of the dielectric.

Returning to the original problem of finding the inductive reactance of an air-dielectric line that is (a) short-circuit-

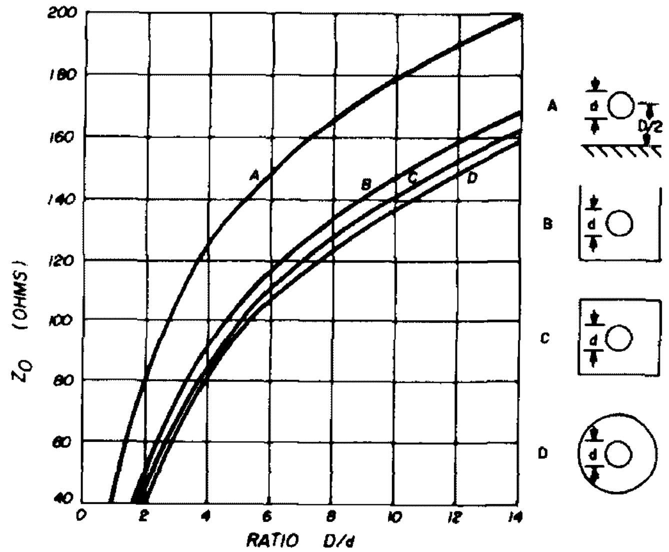



**fig. 2. Transmission-line characteristics using air dielectric. Ratio D/d should be chosen in the range from 2.5 to 5.0.**

ed at the far end, and (b) shorter than ¼ wave, the relationship between inductive reactance and physical length is

$$X_L = Z_o \tan \frac{\text{length in inches x 90}}{\text{length of ¼ wave}}$$

$$= Z_o \tan \frac{\text{length x 90 x freq in MHz}}{2952}$$

$$= Z_o \tan \frac{\text{length x freq in MHz}}{32.8} \quad (2)$$

**Fig. 3** shows the relationship of $X_L$ and length, normalized with respect to $Z_o$ to simplify the curve.

## design procedure

The problem is to assemble all this information into something that can be used to design a tank circuit. The first step is to choose one of the configurations shown in **fig. 2**. The choice will depend on materials available and by the metal work that may be required. For the average workshop, the three- or four-sided square can be built fairly simply from flat metal with angles at the corners.

Next, decide on a suitable ratio of D/d (see **fig. 2**). The recommended values in **fig. 2** were determined from the curves in **fig. 4**, which shows that ratios between about 2.5 and 5 provide lowest attenuation.

Voltage breakdown and power-handling properties are usually no problem for amateur work. (A further discussion on this subject is contained in reference 3, from which the curves in **fig. 4** were obtained.)

From the chosen dimensions, the $Z_o$ of the line must be determined from **fig. 2**. Next, the capacitance that will be present at the tube end of the line must be estimated. This capacitance is composed of tube output capacitance, strays, and the tuning capacitance. Tube plate capacitance

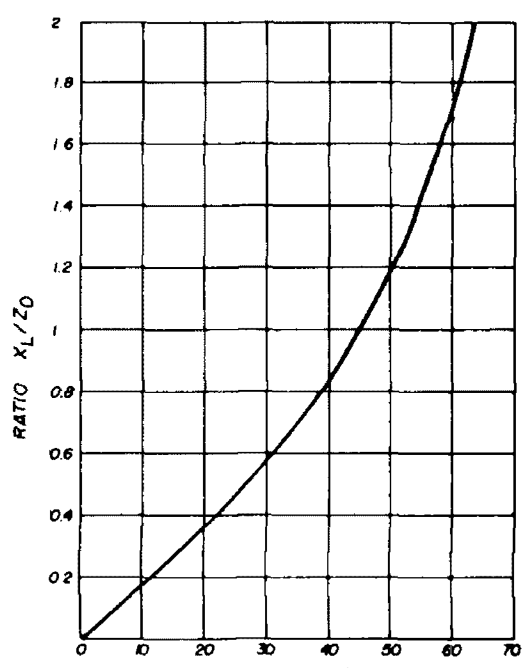



**fig. 3. Relationship between line length and inductive reactance.**

can be obtained from tube-manual data; stray capacitance can be estimated at a few pF; and the tuning capacitance can be the mid-range value of the capacitor, usually just a few pF.

The capacitance at the tube end of the line is the sum of these capacitances. The capacitive reactance, $X_C$, of this sum equals the inductive reactance, $X_L$, required to resonate the circuit at the operating frequency. This value of $X_C$ can be estimated closely enough from **fig. 5**.

After dividing $X_C$ by the line $Z_o$, the line electrical length can be determined from **fig. 3**, and the physical length can be calculated from

$$L = \frac{32.8\,\phi}{f} \tag{3}$$

where
L = line length (in.)
$\phi$ = electrical length (deg)
f = frequency (MHz)

## practical circuit

A means must be provided to feed high voltage to the tube anode. This can be parallel feed through an rf choke, with a plate-blocking capacitor to keep high voltage off the transmission line. However, a better method is to feed the high voltage to the tank through an rf choke at a point on the line where rf voltage is minimum; *i.e.*, at the rf shorted end.

A capacitor is then inserted at the shorted end to complete the rf short circuit on the line and to provide dc isolation of the inner conductor from ground. The capacitor can be built from a metal sandwich separated by a teflon film and installed into the end of the line. The capacitance value should be between 100-500 pF to present very low reactance. The capacitor should have low losses since it must pass high rf circulating currents. Teflon sheet about 10 mils thick is a good choice for the dielectric. The tuning capacitor can be made from two metal discs with variable spacing between them.

The transmission-line center conductor should be as large as possible, consistent with desired $Z_o$, to keep losses down. Silver plating is recommended. (A good

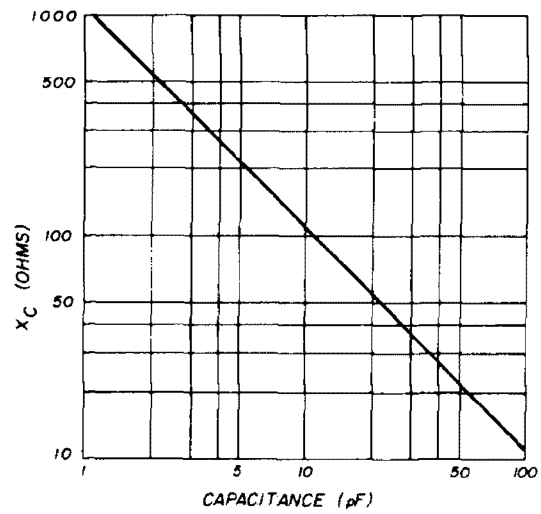



**fig. 5. Capacitive reactance as a function of circuit capacitance at 145 MHz.**

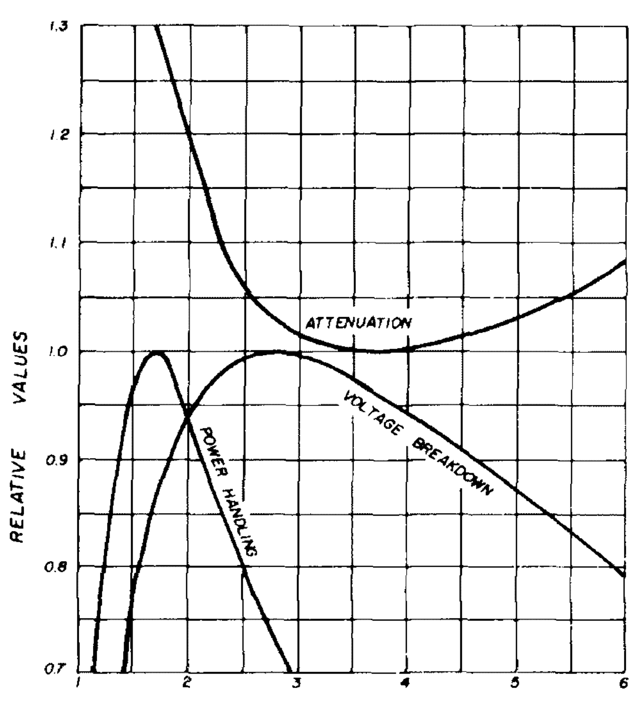



**fig. 4. Line $Z_0$ choices for design.**

*"Cool-Amp," available from Cool-Amp Co., 8621 S.W. 17 Ave., Portland, Oregon 97219. Also, no. 28-203 electroplating set plus 28-211 silver electrolyte; Radio Shack, 730 Commonwealth Ave., Boston, Mass. 02215.

article on silver plating vhf components is contained in reference 4.) Other alternatives are also available.*

Antenna coupling is provided by a coupling loop inserted into the cavity at the cold end, similar to Figure 2-60B in reference 2.

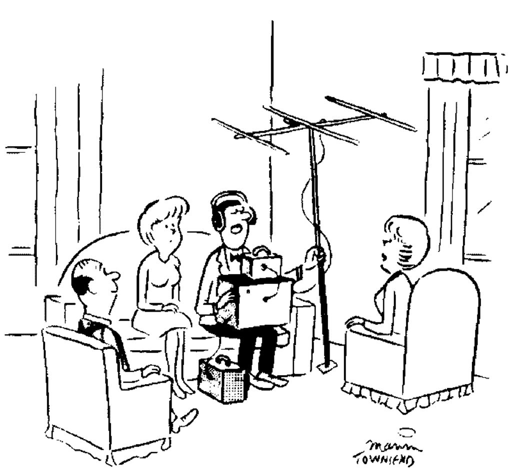
## tank circuit Q

Reference 3 states that a resonant coaxial line optimized for minimum attenuation can have an unloaded Q of about 3000. Since the tank-circuit efficiency is determined by the relationship between unloaded and loaded Q, it's desirable to keep unloaded Q as high as possible. The circuit Q obtainable with a linear tank circuit is much higher than with lumped constants.

## design example

An example will illustrate the principles involved. Choose a 4X150 as an output tube, which has 4.5 pF output capacitance. Estimate stray and tuning capacitances as about 4.5 pF. Total output capacitance is then 9 pF. At 145 MHz, $X_C$ is 120 ohms (**fig. 5**). Now choose a D/d ratio of about 4 and a 3/8-inch diameter inner conductor. This combination gives D = 1.5 inches.

For ease in construction, choose a 4-sided square, 1.5 inches on a side. This type and size of line has a $Z_o$ of 87 ohms. Dividing $X_C$ = 120 ohms by $Z_o$ = 87 ohms yields a ratio of 1.38. From **fig. 3**, the electrical length is 54 degrees. Multiplying 54 by 32.8 and dividing by 145 gives the physical length of the transmission line as 12.2 inches. This is the total path length above the ground plane or chassis and included the path through the tube and the capacitance at the far end of the line.



**"Hope you folks don't mind me bringing my gear . . . I'm expecting an important CQ from Spratley Island tonight."**

## conclusion

My philosophy on building home-brew gear has been that a little basic design should be done first, and then the circuit breadboarded to work out specific details. Even yet I am sometimes pleasantly surprised at the results of a little basic theory applied to practice.

For example, a look at **fig. 2** shows that the four-sided square configuration has a lower characteristic impedance than the three-sided one. Therefore, when a metal top is put on the three-sided line, the characteristic impedance is reduced. It might appear, at first glance, that adding the fourth side would add capacitance to the circuit and hence lower the resonant frequency, but it doesn't work that way. The line resonates with a fixed tube capacitance, so decreasing the line $Z_o$ increases line electrical length to maintain resonance.

Increasing line electrical length is equivalent to increasing the frequency; therefore the resonant frequency becomes higher. The change, though slight, is apparent on a grid-dip oscillator. This example shows that a little basic design saves a lot of cut and try and helps to understand what goes on in these circuits.

### references


1. "Very High Frequency Techniques," Radio Research Laboratory, Harvard University, McGraw-Hill, 1947, Vol. 1, Ch. 15.
2. "Radio Amateurs Handbook," ARRL Staff, 1969, Ch. 13.
3. A. F. Harvey, "Microwave Engineering," Academic Press, 1963, p. 10.
4. R. W. Campbell, W4KAE, "Silver Plating for the Serious Amateur," *ham radio,* December, 1968, p. 62.
5. F. E. Terman, "Electronic and Radio Engineering," McGraw-Hill, 1955, p. 107.


ham radio

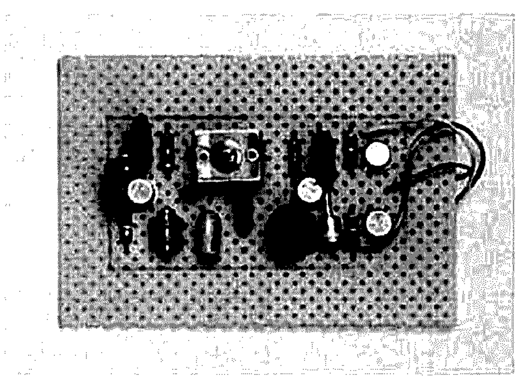
# printed-circuit boards without printing

Roy C. Corderman, W4ZG, 730 Yorkshire Road, Winston-Salem, North Carolina

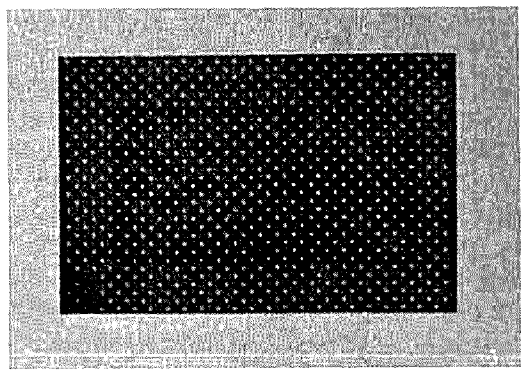
The printed-circuit board is adaptable to quantity manufacture of electronic circuits using small parts operating at low power. For the home workshop, where the quantity is likely to be one board, the conventional process of making the board is tedious and messy. A search was made for an easier and less messy procedure that would result in an attractive product. Here's a report on the result.

## construction

Some circuits can be designed satisfactorily without breadboarding and subsequent modification. However, most circuits designed by home craftsmen require such an approach. So I decided at the start of this project that a working circuit should be made before transferring it to a copper-laminate board.

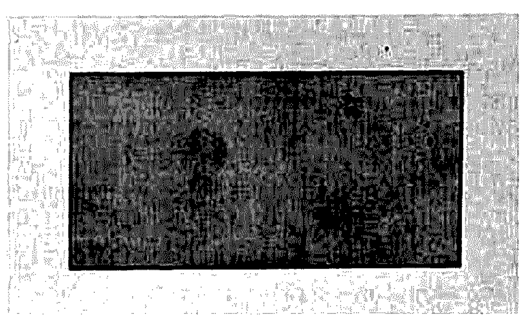


**Step 1.** A diagram of the circuit is first made on paper, using standard drafting aids. When preparing the drawing, use care to keep the number of crossing wires to a minimum. This will save much time when laying out parts.



**Step 2.** Mount the parts on a punched phenolic board (available from electronic parts supply houses) and position the parts as in the drawing you made in step 1. Try to have wires cross underneath parts on the face of the board. Where parts are to be connected, bring the wires together, cut off surplus wire, and solder.

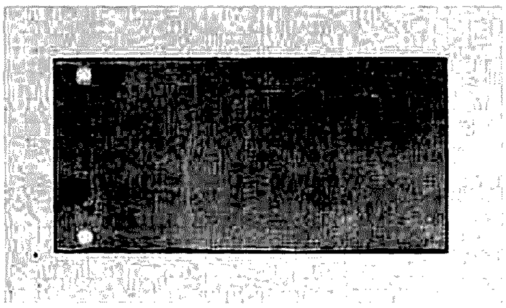
After all parts have been mounted and connected, apply appropriate voltage and check the circuit using temporary terminals and clip leads. Make necessary modifications, then revise the basic drawing.



**Step 3.** Drill any holes to be used for mechanical mounting or for terminals in a

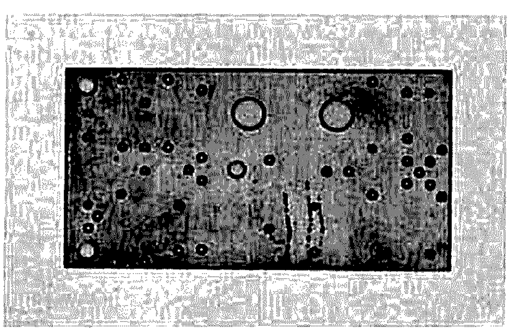
second piece of phenolic board the same size as the board to be used for the finished product. This work is necessary at this point so that final parts layout won't interfere with mounting hardware or terminals.

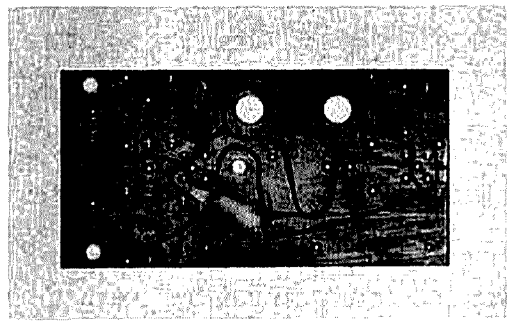
Next, mark the holes on the face of the phenolic board through which leads of parts are to be inserted. Follow the general placement of parts as in step 2, but relocate parts as necessary to provide an even layout and to avoid crowding.



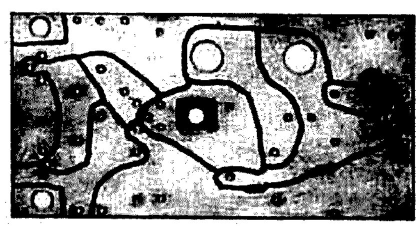
**Step 4.** Cut the copper-clad board to finished size, deburr rough edges, and drill the mounting holes using the phenolic board in step **3** as a template. Hold the two boards together with C-clamps. The marked side of the phenolic board should face up; the copper side of the copper-clad board should face down.

Step 5. Place a piece of hardwood under the copper-clad board. Use a no. 50 drill, and drill through each of the marked holes on the phenolic board where component leads will be inserted. Separate the two boards, then remove any burrs with a larger (hand-held) drill bit. Use fine sandpaper to finish.





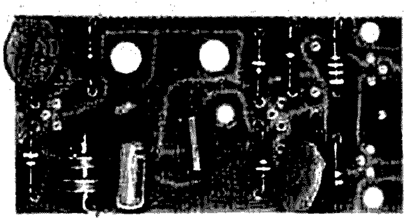
**Step 6.** This step should be accomplished very carefully and work checked before proceeding further. Using the breadboard model on the first phenolic board (step 2), locate on the back the groups of holes connected together. Find the same holes on the back of the copper-clad board, and draw a pencilled line around them. Do this for each of the interconnections until all have been marked on the copper side of the board. Eliminate duplication of lines between connected areas. Wherever possible, extend the lines to the edge of the board. When this has been done, there will be lines on the board where insulating paths are to be cut into the copper.



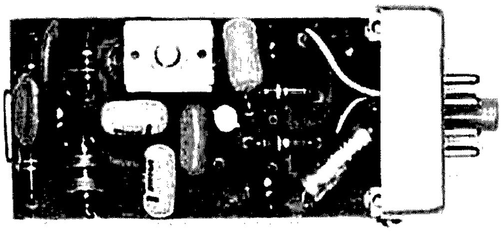
**Step 7.** The next step is to remove the copper from these insulating paths. This can be done with an engraving drill or a dentist's no. 5 or 6 burr drill held in a drill-press chuck.

Place the copper-clad board on a piece of material of uniform thickness, such as a phenolic board about a half-inch thick. Do not use wood; it varies too much in thickness. Tape or otherwise temporarily attach the copper-clad board to the thick phenolic board. Lubricate the underside of the phenolic board so it slides easily on

the drill-press table. Now move the copper-clad board under the drill tip, which should be running at a medium-high speed. Follow the pencil lines marked in step 6. Adjust the depth of cut so all of the copper is removed but little of the insulating material is removed. This operation will be simplified if the insulating paths have been brought to the edge of the board, as suggested in step 6. When all paths have been cut, sand the work to remove any burrs.



Step 8. The next step is to mount the parts. Place leads through holes from the insulated side, keeping appearance in mind as each part is positioned. Bend leads close to the copper on the underside and away from the adjacent insulating path. Cut leads to 1/8-inch on the bent part of the lead. Solder all connections using a low-temperature iron. No soldering guns, please! Use minimum heat on diode and transistor leads. Solder these parts on the insulated island to which they connect. Remove any surplus solder flux with alcohol on a soft cloth.



Mount the board in its container or on its plug-in terminal strip, and connect any wire leads.

## conclusion

This method of preparing PC boards has been used for circuits operating at and above 150 MHz. No interaction problems have been experienced. By properly positioning the various parts, circuits operating at ground potential can be placed between those carrying differing levels of rf energy, thus decoupling them.

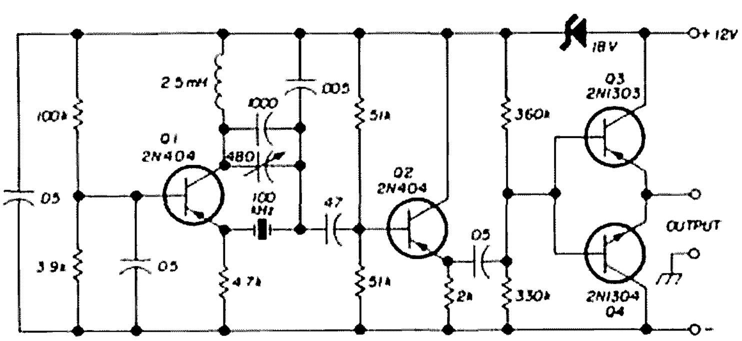



**fig. 1. 100-kHz crystal oscillator used to demonstrate printed-circuit construction. Transistor stage Q2 was found unnecessary and was deleted from the final design.**

The finished product looks nice and works well. The boards are much stronger mechanically than those made with the electrolytic process, since much more copper remains on the board. Changes have been made in the circuits to meet revised needs by using a Dremel hand drill* with a burr drill bit to cut an insulating path through an island and thus separate a circuit.

ham radio

*The Dremel drill is a high-speed 117-Vac unit which operates at 27,000 rpm. Available at hobby shops, or from the Dremel Manufacturing Company, 2420 18th Street, Racine, Wisconsin 53401.

# a simple test set

## for transistors and diodes

Eight components are all you need to build this tester to determine the go, no-go status of unknown devices

Fred Johnson, ZL2AMJ, 15 Byron St., Upper Hutt, New Zealand

Have you ever gazed at a transistor, wondering *if it's an npn or a pnp*? Or have you ever wondered if, in fact, it exhibits any transistor action at all? The test set described here was developed to answer these questions quickly, directly, and with minimum effort. The tester can be used to sort *transistors* into good, bad, pnp, and npn categories. It can also identify good and bad diodes and determine anode or cathode polarity.

### description

Shown in **fig. 1** are the externally visible components: just two lamps, a transistor socket, and a pushbutton switch. Place a transistor into the socket, note that both lamps are out, quickly press the pushbutton switch—the pnp lamp glows, indicating transistor action and a pnp device. The npn lamp would have lit had the device been an npn transistor. Place a diode between C and E pins on the socket, and only one lamp should light. The panel labels (**fig. 1**) identify the cathode end.

### construction

The circuit is shown in **fig. 2**. The 6–8 V supply can be taken from an existing piece of equipment—in fact, the entire unit can be built on a corner of unused panel space in a piece of test gear. It can just as easily be built in a minibox, with a bell transformer for a voltage supply. For these reasons, no construction details are given here. Nothing is critical about construction, layout, or component types.

The resistor in series with the switch should have a lower value if power transistors are to be investigated. A little experimenting will produce the correct value.

### limitations

As with any piece of simple test gear, this one has some limitations. However, as a quick tester, it has no equal. The data in **table 1** can be used to interpret the test results.

The collector current required for lamps may be excessive for some transis-

tor types and could cause the device under test to heat faster than expected. Prolonged operation may cause transistor failure. No failures are likely if the switch is operated just long enough to identify lamp operation. The prototype unit has been in operation for a long time, and so far as I know, it hasn't been responsible for any device failures.

table 1. Interpretation of test results

| device under test | switch | pnp lamp | npn lamp | conclusion |
|---|---|---|---|---|
| diode between E and C pins | — | on | off | good unit; anode connected to E pin |
| | — | off | on | good unit; anode connected to C pin |
| | — | off | off | unit open-circuit |
| | — | on | on | unit short-circuit |
| transistor: emitter to E, base to B, collector to C | — | on | on | short between collector & emitter |
| | — | off* | off* | normal—i.e. good unit |
| | on | on** | off | good pnp unit |
| | on | off | on** | good npn unit |
| | on | off | off | open-circuit collector or a good small-signal unit with low gain |

* Some modern (planar) transistors exhibit a breakdown when a reverse potential is applied to the collector with the base circuit open. This may show up as a faint glimmer on one lamp during this test. This is normal, and the other lamp should glow brightly during the next test when the switch is pressed.

** A dim lamp during this test usually indicates that the transistor has a high saturation voltage and should probably be rejected.

## how it works

The signal applied between base and emitter with the switch pressed is an ac signal; i.e., terminal B voltage changes positively and negatively with respect to the E terminal. A transistor will conduct only when its base-emitter junction is forward biased; positive for npn and negative for pnp. So when B is positive, forward biasing the npn base-emitter junction, collector current will flow through the npn lamp. The pnp lamp will not light, because its series diode is reverse biased and nonconducting.

On the next half cycle, the base-emitter junction is reverse biased and the transistor is cut off. The transistor therefore only conducts on alternate half cycles, and the lamps indicate accordingly.

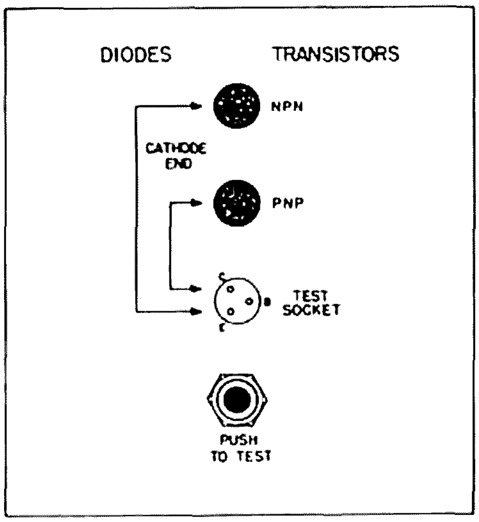



**fig. 1. Suggested panel layout for the tester.**

A short circuit inside the transistor between emitter and collector will result in both lamps lighting. An open-circuit collector lead will result in neither lamp lighting. With the switch released, the transistor acts as two diodes in series opposing; hence neither lamp will light. When the switch is pressed, transistor action operates a lamp.

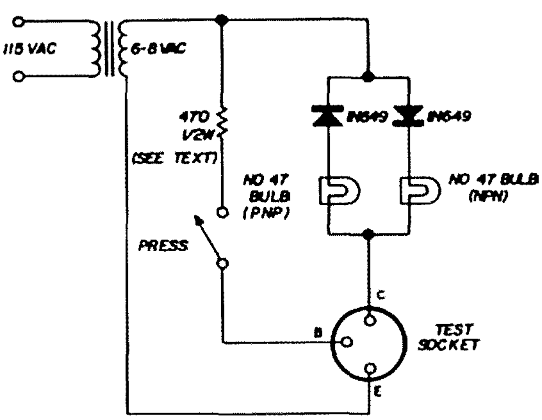
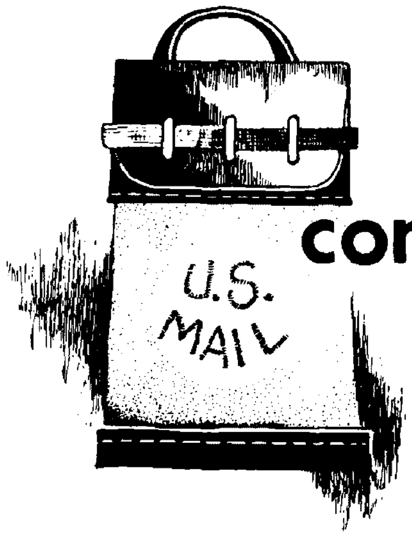



**fig. 2. Schematic for the semiconductor device test set.**

A diode connected between the E and C pins will conduct on one-half cycle only and only one lamp will light unless it's short-circuited, in which case both lamps will light. The diode connections can be identified from the appropriate lamp indication.

ham radio




# comments

## reflected power

**Dear HR:**

I was very interested in VE3AAZ's very good paper in the May 1970 issue of *ham radio,* "Some reflections on reflected power." In the section on line reflections he asks, "What happens if the load impedance doesn't equal the line impedance and some energy is reflected?" Then he goes on to answer his own question. "It's unlikely that the source and line impedances will be exactly equal. Thus, any energy reflected from the load will travel down the line to the source to be reflected again toward the load. This repeats until the wave's amplitude becomes too small to be of interest."

Thus his concept of the reflected wave being re-reflected from the junction between the source and line depends upon the source and line impedances being unequal. But if this is correct, then one must envisage various situations in which those two impedances differ in various ratios, in which cases one would expect various proportions of the reflected wave to be re-reflected on getting back to the source-line junction. Thus it may be incorrect to say that *"any* energy reflected from the load will travel down the line to the source to be reflected again toward the load." (That is to say, any departure from equality of the two impedances at the sending end will cause complete reflection.) But suppose, in the rare case, that these two impedances are exactly equal; will there be no reflection? This seems to be the conclusion to which we are driven by adopting Mr. Anderson's concept which, by the way, has wide acceptance though seldom specifically stated.

However, I do not believe that this is the actual situation. To explain complete re-reflection at the sending end junction, appears to require that the oncoming reflected wave see either zero impedance or infinite impedance as those are the only conditions for complete reflection. What the whole explanation may be I do not know. It may well be that the hypothetical physical model invoked in nearly all discussions of transmission lines differs from the mathematical model sufficiently to be inadequate to explain reflections at the sending end. I would like to see this point discussed in *ham radio.*

**Hubert Woods**
**Guadalajara, Mexico**

*The statement that "any energy reflected from the load will travel up the line . . ." means simply (in a lossless line) that all the energy reflected from the load goes back to the source end where something happens to it. It does not imply that the energy reflected from the load will go back to the source and then all come back toward the load again.*

*My original manuscript was in two parts which was condensed for publication. The original said, "And, if we have reflected energy traveling back to the source, then what? When the line looks back into the filter/transmitter, there is scarcely more than one chance in a million that it will see an impedance equal to $Z_O$, so at least part of the reflected energy (which cannot simply disappear according to the law of conservation of energy) is reflected again and starts toward the load where it is again reflected (in part) and what's left of it goes back to the source and so on and on, back and forth, until the wave gets too small to interest us... One group (of waves) is moving source-to-load and is called the "forward" or "incident" component: The other group is moving load-to-source and so is called the "reflected" component."*

*In the interests of brevity I avoided any reference to reflection coefficient in my original paper. This quantity is defined*

$$\rho = \frac{Z - Zo}{Z + Zo}$$

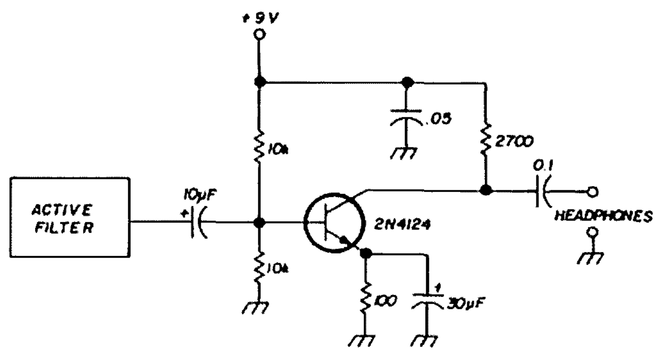
*The customary procedure is to make $Z = Z_L$ (the impedance of the load). However, if one elects to use the "successive reflections" method to find out what is happening on the line, then one must define an additional $\rho$ in which $Z = Z_S$ (the impedance of the source). In any event it is possible, though tedious, to use these coefficients to completely solve for the voltages, currents and power on the line. There is clearly just as much variety possible in the sending end $\rho$ as there is in the load end $\rho$. But if the load $\rho$ is zero (swr exactly one) there will be no reflected energy and the sending end $\rho$ is only of academic interest. The important point to note is that the load end gets the first "crack" at the energy and so determines how much or how little energy there will be for the sending end to work on. Accordingly, the swr depends upon the load $\rho$ only.*

Walter Anderson, VE3AAZ
Toronto, Canada

## active audio filter

Dear HR:

I wanted to let you know that I've constructed the variable bandpass audio filter that you ran in the April issue of *ham radio.* I use a 2N4124 in the filter, and another 2N4124 as a single audio stage for headphone operation. Together, they produce a quite usable level of audio with really great selectivity. The amplifier stage is very simple:




I am in the process of using this filter and audio as a replacement for the filter in the DC-80-10 receiver of *QST,* April 1969. Without exaggerating, the receiver and this filter produce *better* selectivity than I get with my RME-4350 with the crystal filter notched all the way in! With the filter at the point of near-oscillation, S-4 signals will come up to a level equal with an S-9 plus 10-dB signal only 100 cycles or so away. I'm amazed at it! Guess I'll never build that transistor version of the old 'selecto-ject' you ran a few months back.

I found the main problem was lack of skirt selectivity—the bandpass signal was really peaked, but signals off the sides suffered little attenuation. Solution: I cascaded two sections of the filter with some minor modifications to adapt to the cascading. These consisted of eliminating the resistor between input and ground in the second stage, and adding a trimmer-type potentiometer to the first stage instead of the regular potentiometer. With this setup, the first stage is adjusted not quite at minimum bandpass (backed off slightly to the point where the stage

still yields worthwhile gain), and the second section is used to adjust bandpass width and depth. Now the signals almost totally disappear above 1100 Hz—they just aren't there anymore!

One thing I've noticed about the filter that is worth mentioning: the input to the filter must be at an adequate level to use its full potential selectivity. I got the impression from the article that the lower the input, the better.

Ade Weiss, K8EEG/Ø
Meckling, South Dakota

## filters for speech clipping

Dear HR:

A good source of 9-MHz filters for use in speech clippers is Spectrum International. I contacted them and learned that their KVG crystal filters were 9 MHz± 200 Hz — off the shelf. On receiving two filters for a transistorized rf clipper project, the center frequencies were within 60 Hz. According to my spectrum analyzer this is a usable tolerance.

Wayne W. Cooper, K4ZZV
Miami Shores, Florida

*For more information on KVG crystal filters, see the report on page 86 of the April, 1970 issue of* ham radio. *The filters may be purchased from Spectrum International, Post Office Box 87, Topsfield, Massachusetts 01983.*

## light-emitting diodes

Dear HR:

A number of new visible-red LED's have been put on the market, including the Hewlett-Packard HP5082-4403 for \$2.50, and the Motorola MLED600 for \$2.25 (single-unit prices). Anyone experimenting with these devices should get a copy of "Solid-state Lamps—Part II Applications Manual" from the Miniature Lamp Department of General Electric in Schenectady, New York.

Martin Davidoff, K2UBC
Syracuse, New York

75 cents

focus on communications technology . . .

# ham radio

magazine

DECEMBER, 1970

## this month

December, 1970
volume 3, number 12

**staff**

James R. Fisk, W1DTY
editor

Nicholas D. Skeer, K1PSR
vhf editor
J. Jay O'Brien, W6GDO
fm editor

Alfred Wilson, W6NIF
James A. Harvey, WA6IAK
associate editors

Jean Frey
*art director*
Wayne T. Pierce, K3SUK
cover

T. H. Tenney, Jr. W1NLB
publisher

John B. Morgan, K1RA
advertising manager

**offices**

Greenville, New Hampshire 03048
Telephone: 603-878-1441

**ham radio** *magazine* is
published monthly by
Communications Technology Inc.
Greenville, New Hampshire 03048

Subscription rates, world wide:
one year, $6.00, three years, $12.00.
Second class postage
paid at Greenville, N. H. 03048
and at additional mailing offices

Foreign subscription agents:
United Kingdom:
Radio Society of Great Britain,
35 Doughty Street, London WC1, England.

All European countries:
Eskil Persson, SM5CJP, Frotunagrand 1,
19400 Upplands Vasby, Sweden.

African continent:
*Holland Radio*, 143 Greenway,
Greenside, Johannesburg,
Republic of South Africa

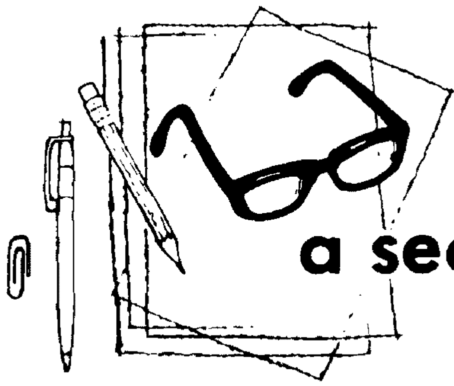

Printed by Wellesley Press, Inc.
Wellesley, Massachusetts 02181, U.S.A.

**ham radio** is available to the blind
and physically handicapped on magnetic tape
from Science for the Blind,
221 Rock Hill Road, Bala Cynwyd,
Pennsylvania 19440.
Microfilm copies of current
and back issues are available
from University Microfilms,
Ann Arbor, Michigan 48103.

Postmaster: Please send form 3579 to
**ham radio** magazine, Greenville,
New Hampshire
03048

**ham radio** *magazine*

# contents



# a second look

by jim fisk

**Regular readers of this page** may recall an editorial in the April 1970 issue by Skip Tenney, W1NLB, publisher of *ham radio.* The central theme of Skip's message concerned the growth of amateur radio and the attitude of hams in contributing to this growth – not only in quantity of numbers, but in quality of technical achievement. Such balanced growth is clearly essential if ham radio is to survive the challenge of the technology explosion in electronics.

One subject Skip touched upon was the novice ham and his problems in becoming established as a permanent member of the amateur radio community. It was mentioned that about 50% of those passing the novice test never get on the air. Carrying this thought a little further, of those who do get on the air perhaps less than 30% develop the skills and knowledge necessary to progress to higher license grades.

There are many reasons, of course, for this poor success ratio. One is plain discouragement. Novices need the help and mature direction from those who have been through the mill. Many novices build or buy equipment but have only a vague idea of how the equipment really operates beyond basic tuning adjustments. Experienced hams, who are sincerely concerned about the development of amateur radio, can help these novices over the rough spots of their transition period. Each novice who does manage to progress to general-class status is a potential contributor to the common cause of ham radio.

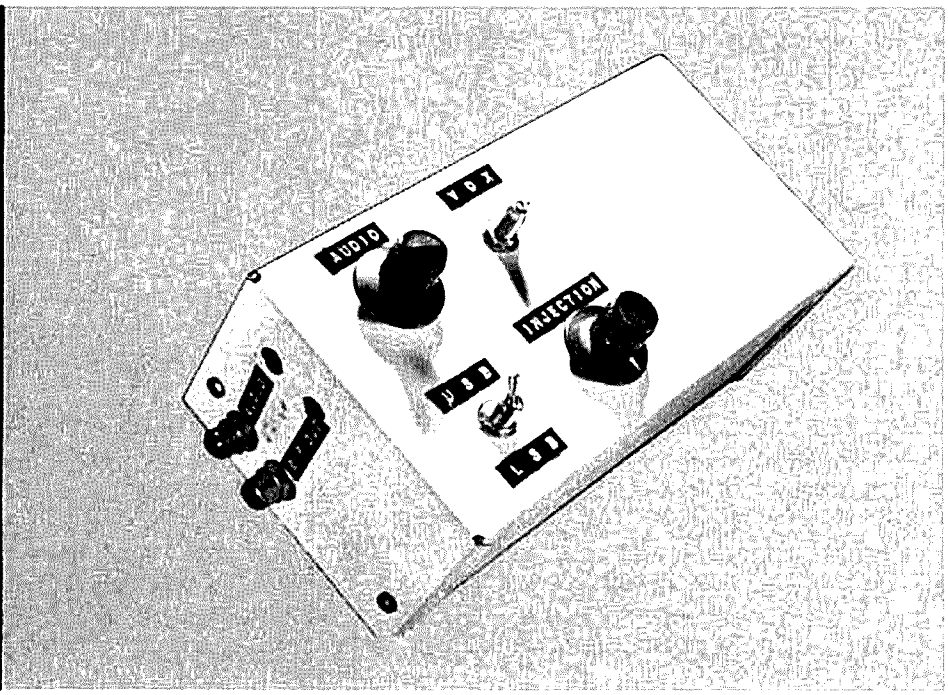
Tune across the novice frequencies on any band during the evening hours, and you'll hear hundreds of novices endlessly calling CQ with negative results. It is any wonder so many become discouraged?

One way the experienced ham can help the situation is to make contact with a few novice stations and offer technical aid if it seems appropriate. Some novices object to general and higher-class hams invading novice territory on the grounds that these bands are already overcrowded. However, it certainly seems logical that any aid given a novice by a more experienced operator would compensate for the additional band occupancy.

One example recently brought to my attention bears this out. An extra-class ham contacted a novice who was having trouble with his antenna, which in turn was causing transmitter loading problems. Actually, the novice solved the problem himself after a few friendly words of advice. All their activity consumed the better part of two hours at 12 wpm. In addition to the technical help, the contact provided some needed code practice for the novice ham, and the extra-class ham obtained some practice in forbearance.

If you've become bored with the ordinary ham operating activities, why not tune across the novice frequencies and see what's happening? A few hours a week exploring the world of the novice and his problems can be rewarding if you enjoy helping people. Who knows? Maybe your efforts might provide the amount of impetus necessary to make the difference between a dropout and a winner.

**Jim Fisk, W1DTY**
**editor**




# a filter-type ssb generator

This construction project features a basic ssb generator circuit that can be used in hf or vhf exciters

R. Bain, W9KIT, 4915 Ridgedale Drive, Fort Wayne, Indiana

The state of the art being what it is today, the homebrewer seems to be a vanishing breed. For those who cut their teeth on tubes, solid state is somewhat perplexing. But then, even the ham-equipment manufacturers, with a few exceptions, have been somewhat reluctant to plunge into wholesale change-over from tubes. Single sideband also has popped up to further complicate the homebrewer's life. Though a delight to use, it's another matter when it comes to generating ssb signals. There are still those hardy few who hold that building is half the fun of getting there. For these, here is a 9-MHz selectable-sideband filter-type ssb generator.

## description

Referring to fig. 1, two stages of audio amplification increase the microphone output to the proper level to drive the balanced modulator. A switchable 9-MHz

oscillator drives an emitter-follower isolation stage, which in turn provides both drive to the balanced modulator and a variable-level carrier for injection into the output amplifier. An additional audio stage and a rectifier/filter network provide a dc vox voltage.

## audio stages

A schematic of the complete ssb generator appears in fig. 2. The first audio stage, Q1, is a 3N128. It provides the high input impedance suitable for a ceramic microphone. The rf choke and bypass capacitor in the input help prevent any rf feedback. The first stage provides about 20 dB gain.

The second stage, Q2, provides about 30 dB gain or more and has a low-pass network in its output to restrict the high-frequency audio response. This network has the dual purpose of improving the supression of the audio components above 3 kHz and eliminating any rf at this point. There is also some low-frequency rolloff below 300 Hz, which is provided by small values of coupling and bypass capacitors in the first stage. This low-frequency rolloff helps suppress the opposite sideband components near the carrier frequency and eliminates the non-essential low-frequency components in the wanted sideband.

The third audio stage provides a low-impedance output to the balanced modulator. This circuit has a pair of diodes connected as a limiter, which allows about 8 dB compression without objectionable distortion. The level at this point is 0.7 Vrms maximum.

An additional output with variable level is provided from Q3 to drive Q4, the vox amplifier. The two diodes and filter network driven by Q4 provide a maximum of about 6 Vdc to drive an external vox-control relay. This voltage can be made poistive by reversing the diodes.

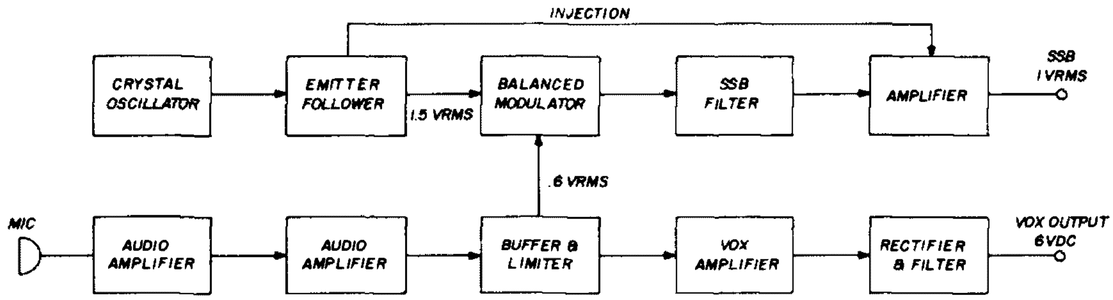



**fig. 1. Block diagram of the 9-MHz selectable-sideband ssb generator.**

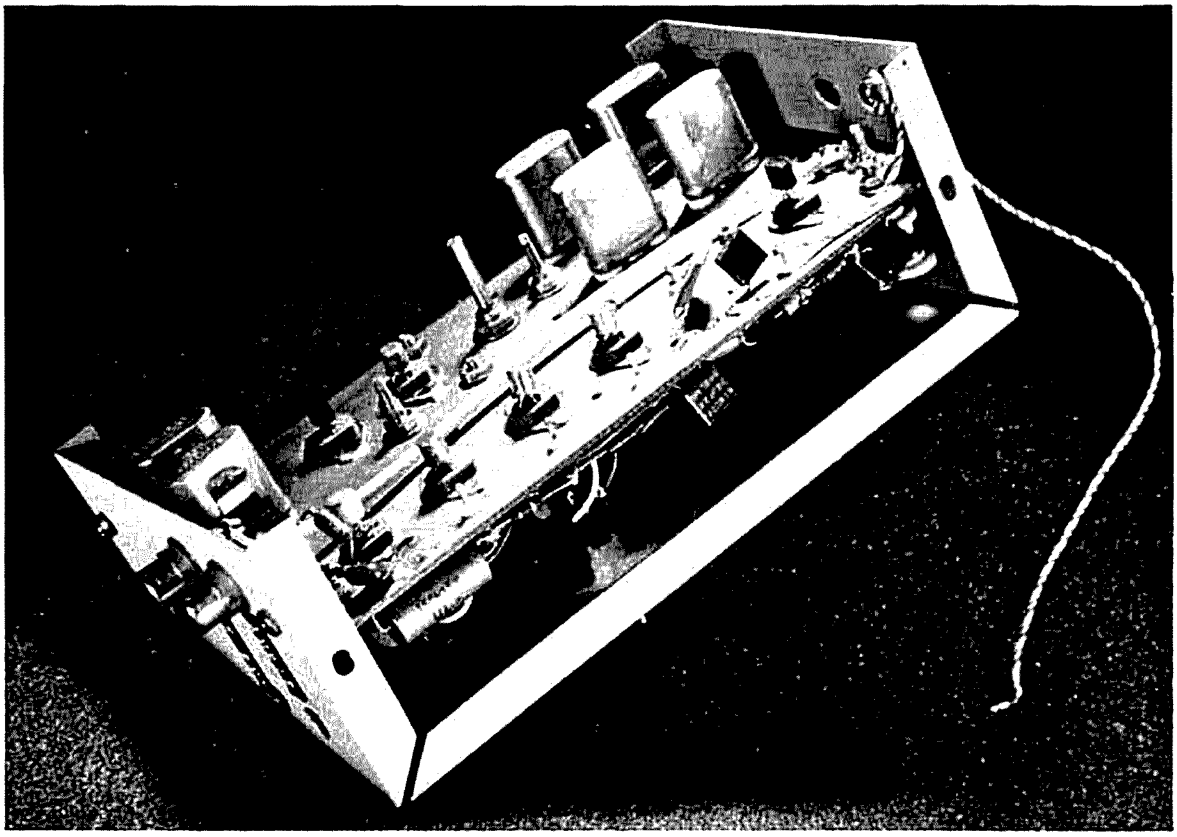
## carrier oscillator

The carrier oscillator, Q5, provides a carrier on either the upper or lower slope of the filter bandpass to produce lower and upper sidebands. The carriers ideally should be at the -6 dB points or lower. The capacitors across the crystals locate the carriers on the skirts.

The emitter-follower stage, Q6, provides isolation between oscillator and balanced modulator. The pot in the emitter of Q6 provides a variable-level carrier for tune-up, cw, and a-m operation. The input coupling network to the follower provides both bias and the proper drive level. If a 1-pF capacitor isn't available, use a capacitive divider to keep the drive level down.

## balanced modulator

The balanced modulator is of the two-diode variety, with unbalanced inputs for the audio and rf, and balanced output. The two diodes should be reasonably matched for forward resistance. The rf level should be about 1.5 Vrms into the arm of the balance pot. The ratio of rf to audio should be greater than 2 to 1. Since the limiter in the audio circuitry holds the audio to about 0.6 Vrms, this condition is met. The diodes couple into a balanced bifilar winding (the primary of T1). A bifilar winding is shown in fig. 3.



**General arrangement of parts. Chassis is 0.032-inch double-clad board; unit is housed in a box fashioned from aluminum siding.**

## crystal filter

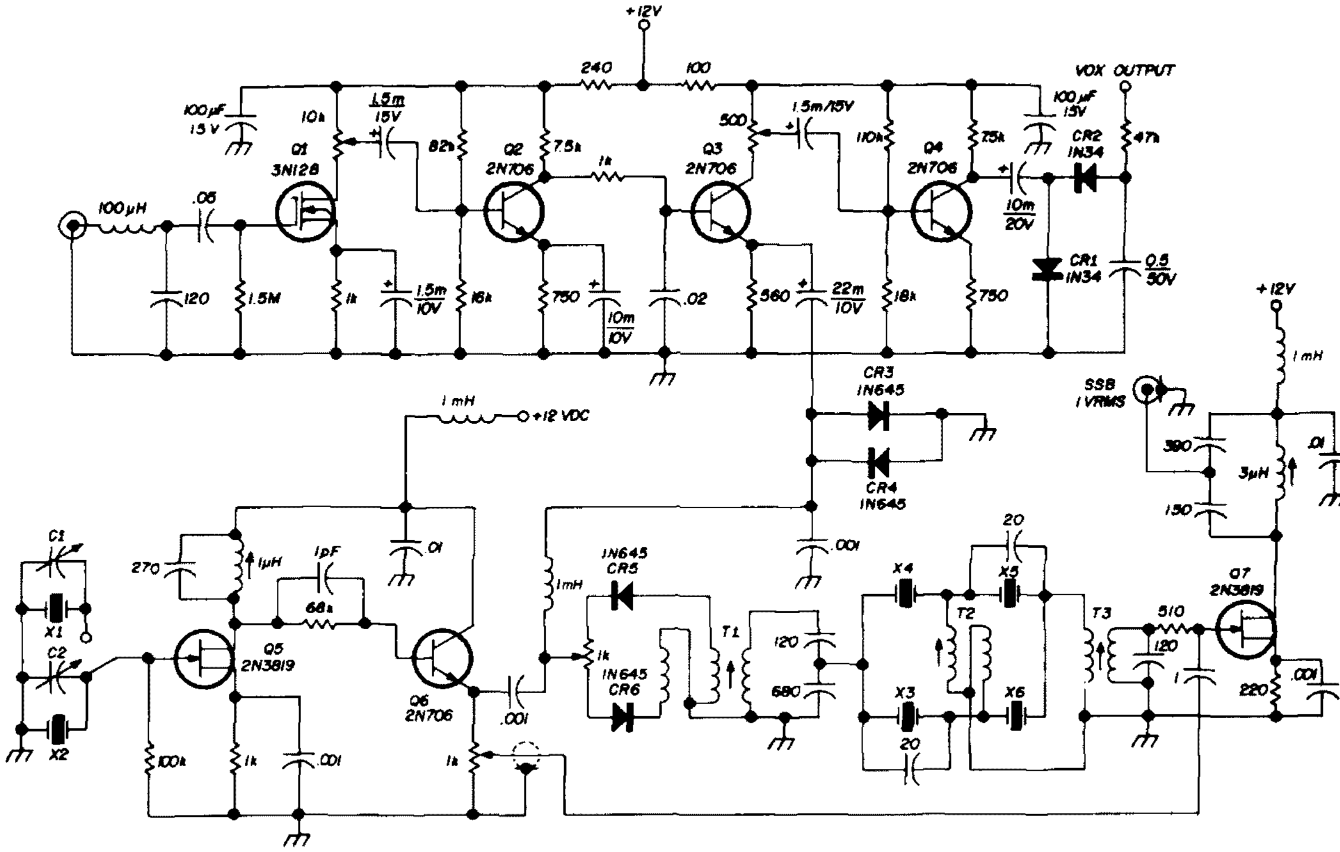
Crystal-filter design may seem difficult, but I have built two of this type with satisfactory results. The input to the filter is stepped down in impedance by a capacitive divider across the balanced modulator output transformer. The crystals used in filter were ordered in HC6 holders, series resonant, 2 kHz apart. Filter bandwidth is about 2.2 kHz at the 3-dB points. Transformer T2 is not tuned; the core is centered in the bifilar winding. The output of the filter is stepped up in impedance by T3 to provide drive to the gate of the output stage. You might think about buying a commercial filter if you don't want to bother with the alignment problems, although a filter can be built for less than half the price of a commercial model.

The output amplifier, Q7, increases the filter output to 1.0 Vrms through a capacitive divider into a 1k load. The 510-ohm resistor in series with the gate was needed to stabilize this stage. Lower impedance loads can be fed by changing the ratio of the capacitors in the divider.

## construction

The ssb generator is constructed on .032-inch double-clad board, using transistor sockets and standoff terminals as required. It would be easy to use an etched-circuit board. Having a copper layer on the transistor side of the board provides good shielding and a good ground plane to prevent ground loops. An inch-high shield is installed between the audio and output amplifier stages on one side of the board and the rf circuitry on the other side of the board. Another shield separates the oscillator from the balanced modulator.

The component side of the board faces the front panel of the unit. Short wires are run from the board to the controls on the front panel. The control location and board layout were coordinated to place the controls over the circuits to which they connect. A ground was necessary between the top of the balanced modulator shield and the front panel to prevent rf leakage from the oscillator to the balanced modulator. The oscillator




C1, C2 2 — 20 pF (variable or selected)

CR5, CR6 matched for forward resistance

T1 4 turns bifilar primary, 3 uH secondary*

T2 6 turns bifilar wound on 1/4" form

T3 5:1 ratio, 3 uH secondary*

X1 8.9995 MHz (parallel resonant with 32 pF)

X2 9.0015 MHz (parallel resonant with 32 pF)

X3, X5 9.0020 MHz (series resonant)

X4, X6 9.0000 MHz (series resonant)

*3 uH coils are J. W. Miller 40A336CBI

**fig. 2. Ssb generator schematic. The circuit includes provisions for af-component conditioning and speech compression. A voltage is also available for an external vox relay.**

crystal cans are grounded by clips on top of the crystal sockets.

The generator is housed in a 6-1/8 x 3-1/4 x 2-1/2-inch homebrew

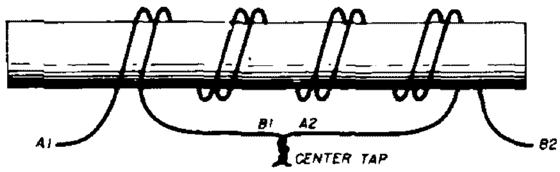



**fig. 3. Bifilar winding used on transformers in the balanced modulator and filter sections. A1 indicates the beginning of one wire; A2 is its other end. Wire B is wound next to wire A. Wires should be twisted for best balance.**

minibox made from a piece of aluminum siding. The board is held in place by no. 4 screws through brass tabs soldered to the corners of the board. In my application, the box is mounted behind the front panel of a homebrew exciter, with the controls projecting through the panel. The controls are fastened to the front panel of the box with silastic so the controls will stay in place when the nuts are removed.

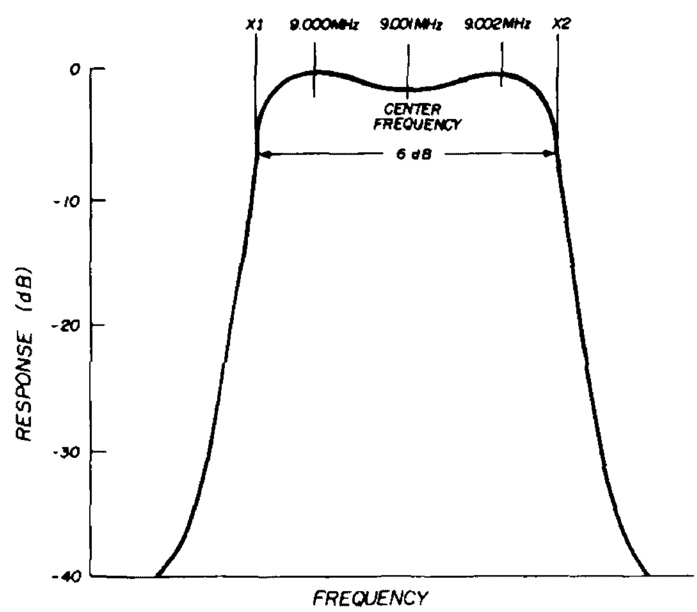



**fig. 4. Crystal-filter characteristic. X1, X2 show carrier placement; 9.000 and 9.002 MHz represent filter bandpass.**

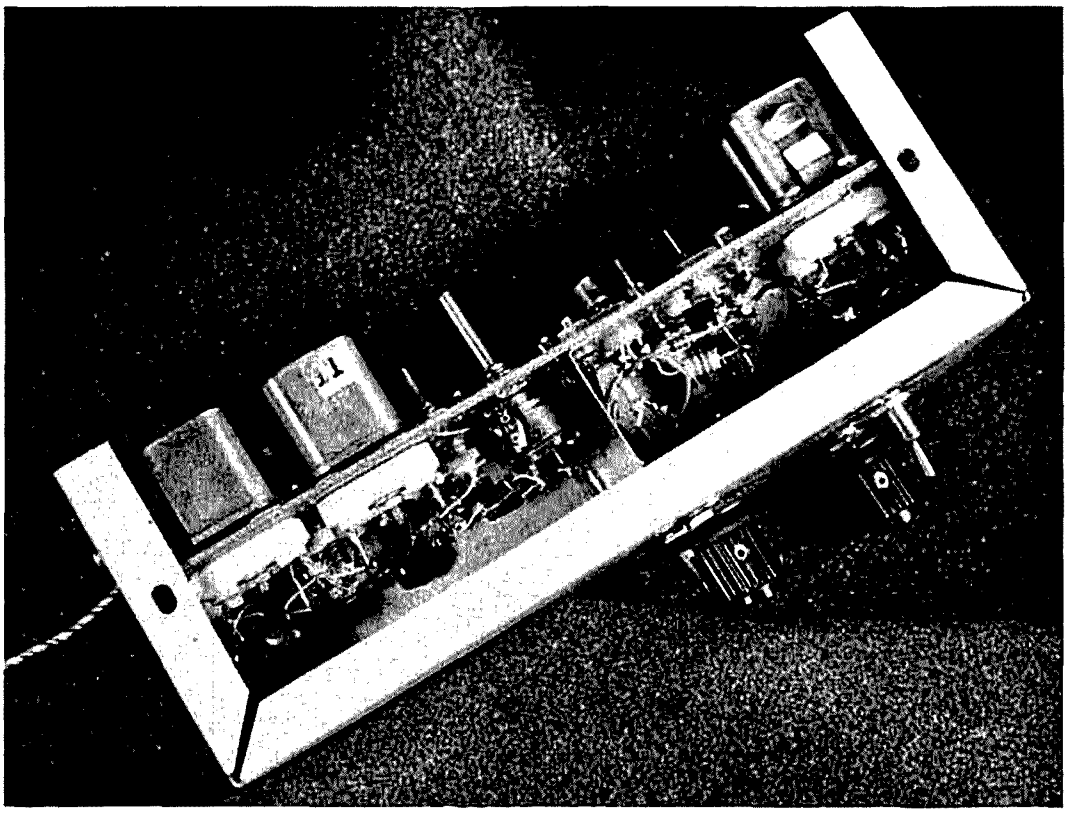
## alignment

The tuned circuit in the oscillator output should be set far enough above the operating frequency to ensure starting and equal outputs with either crystal. The capacitors across the crystals should be adjusted so the carriers fall on the -6 dB points on either side of the crystal filter bandpass. See **fig. 4** for carrier placement. The crystals in the oscillator should be a bit low in frequency to operate with the standard 32-pF capacitance in the parallel-resonant mode. Frequencies of 8.999 and 9.001 MHz will do. The lower-frequency crystals may be easily pulled up in frequency by reducing the capacity across them.

When properly aligned, the filter bandpass should be as in **fig. 4**. Several methods can be used to align the filter. The most desirable method is to inject an fm signal into the filter and display the output on a scope synced to the fm sweep. In the absence of this test equipment, an audio signal can be injected into the rf choke feeding the balanced modulator pot. If the amplifier output is observed on a receiver S meter, the filter bandpass can be plotted as the frequency of the injected audio is swept from 100 Hz to 4 kHz. The filter bandpass char-

acteristic is affected by the tuning of T1 and should be adjusted for equal level of the two peaks. Capacitor C1 can then be adjusted by setting it for minimum capacitance, watching the level on the s-meter, and increasing C1 until about 6 dB (or one S unit) decrease is noted. Capacitor C2 is set to maximum capacity, then backed out for the same results as with C1.

(together with a crystal oscillator), and this mixer drives a second mixer together with a 5.0 to 5.5-MHz vfo for 80 – 10 meter output. The first mixer can be omitted, and output can be obtained on 80 and 20 meters only (9 MHz plus or minus the vfo frequency). For 6-meter operation, a vxo could be multiplied to 41 MHz and mixed with the 9-MHz ssb signal to obtain 50-MHz output.



**Another view of the ssb generator. Shielding between compartments is important to eliminate crosstalk between rf and audio stages.**

The balance pot should null the carrier to −35 or −40 dB. The additional −6 dB from the filter will make the carrier rejection a minimum of −40 dB. If this can't be achieved, the carrier may be leaking around the balanced modulator, or a small balance capacitor may be needed on one side of the balance pot.

## applications

Numerous combinations can be used for vhf or hf ssb transmitters. In my transmitter the generator drives a mixer

The bandwidth of the ssb filter may be a bit narrow for some voices, but it can be easily increased by using 2.5 or 3.0 kHz spacing on the filter crystals. I've had no complaints about the intelligibility, and I am able to compete for the dx with those who use exciters with similar power levels.

Why not try a little construction? There's nothing to compare with the satisfaction gained in creating something worthwhile.

**ham radio**


# radio-frequency interference

Man-made interference can raise havoc with radio communications. Here's how to locate and cure troublesome noise sources.

Jim Fisk, W1DTY

As the number of sunspots goes down, and solar activity declines, more and more amateurs will be emigrating from 10 and 15 meters and lining up for clear spots on 80 and 40. This year there should still be a few good openings on the upper h-f bands, but good DX propagation during the evening hours will probably be limited to 80 and 40, and occasionally, 20 meters.

If you haven't had your receiver tuned up on 80 or 40 recently, you've probably forgotten the man-made interference you run into on our lower bands – hash from 15 kHz horizontal oscillators in local television sets, noise from vacuum cleaners, washing machines, refrigerators and other electric-motor appliances, as well as interference from furnace ignitors, fluorescent lights, and thermostats.

If you live in the country the problem is not as severe as in the city, but the ambient noise level on a cold winter night can still cover up a weak signal on the low end of 80. If you live in the city, you're apt to turn on the receiver, scan the band, note the S8 noise *level*, and opt for a quiet evening of television.

Unfortunately, most of the man-made noise that interferes with amateur communications must be cured at the source. If you put a brute-force line filter (see **fig. 1**) in the ac power line to your receiver, install a noise blanker, and make sure everything is well grounded you've taken steps in the right direction, but you can still look forward to a few quiet evenings in front of the one-eyed monster; noise blankers simply aren't effective against all

forms of noise, and some of the noise is picked up by the antenna, not the line cord.

## tracking the noise

The first problem is to find the source. After you find the culprit you can take the necessary steps to suppress the noise. If you have a mobile setup, you can cruise around the neighborhood looking for the area where the noise is the loudest. However, this approach works best in the country where the houses aren't too close together. If the noise source comes from a densely populated residential area or a high-rise apartment building, your mobile receiver isn't portable enough to track down the source.

The best bet for tracking down noise is a small lightweight battery-powered receiver that is easily carried about. The late Davco DR-30 is ideal, but not too many of us are lucky enough to have one in the shack. A simple direct-conversion receiver such as the one described by K1BQT[1] is perfectly suitable, as are any number of other simple receivers. The other alternative is to build an inexpensive receiver expressly for the purpose of locating troublesome noise sources. An excellent low-cost unit that tunes from the broadcast band to 225 MHz was described by W1CER in *QST.*[2]

## finding the source

Tune your noise-seeking receiver to the highest frequency where you can hear the interference, back down on the gain until you can just hear the noise, and start driving. If the noise drops out, turn around and go in the opposite direction. The noise level will build up as you approach the source. Back off on the rf gain and keep driving until the noise disappears again. Turn around then go back, carefully checking the noise level. After two or three tries you should be able to pinpoint the source within two or three houses. Now get out of the car and walk up and down the street until you can determine which house is creating the interference.

Once you have located the noise source, you can't go any further until you talk to the householder and get his aid in tracking down the exact source. Be diplomatic. Don't accuse him of interfering with your receiver; ask him if he is having interference to his television or a-m broadcast set. If the noise source is in his house the interference will be more severe and he will probably ask you where it's coming from.

Turn up the gain on your receiver so he can hear the interference and show him that it's the same on his tv set or broadcast receiver. Then explain that you'll have to go through a process of elimination to locate the problem. Ask him to turn off his circuit breakers one at a time to make sure his house is *not* the source. *If* the noise disappears with the click of one of the circuit breakers, check out all the appliances in that circuit until you find the offending device. Then ask him to have it corrected — there are filters on the market for practically every type of electrical noisemaker.

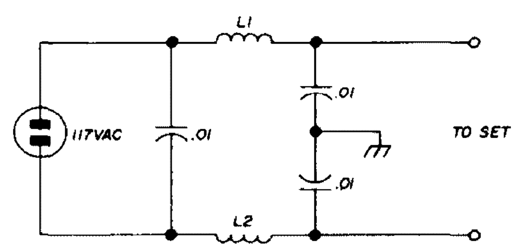



**fig. 1. Brute-force power-line filter. L1 and L2 are 28 turns no. 14 enameled, ½" diameter, spaced to a length of 3½" (use a ½" wooden dowel for a form). The two coils should be wound in opposite directions.**

On the other hand, if the interference isn't coming from his house, you have more detective work to do. When you have found the source and had it corrected, let him know the problem has been found, and thank him for his assistance. He'll be grateful for your help in getting rid of the interference and will probably tell his neighbors — score one for amateur radio.

For more information on identifying and tracking down radio interference, read the excellent articles by WA6FQG.[3]

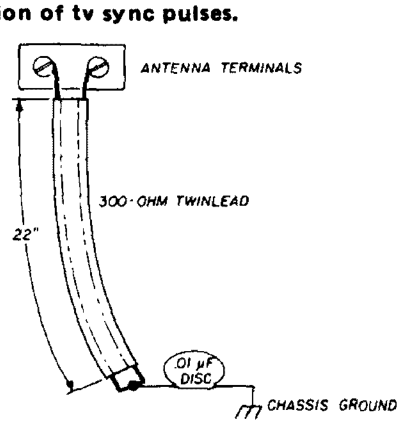
## tv sync interference

*One of the more* troublesome sources of radio interference is the 15,750 Hz horizontal sweep oscillator found in every television set. The fast rise time of the output waveshape results in harmonics that can often be heard on the 28-MHz band. The FCC has radiation-suppression requirements for new sets, but these can still cause interference on 80 and 160 meters. Older sets have no suppression and have been known to cause interference as high as 28 MHz! Tv sync interference is easily identified because it takes the form of rather unstable ac-modulated signals spaced at intervals of 15.75 kHz.

Tv sync interference is usually radiated by the line cord, the antenna feedline or the sweep circuit wiring. In many cases it can be cured by installing a highpass filter at the antenna terminals and bypassing each side of the ac line where it enters the tv chassis with .01 μF disc capacitors.

If these two steps don't stop the interference, cut a piece of 300-ohm twinlead 25 inches long and strip 1½ inches of insulation from each end as shown in fig. 2. Twist the leads of one end together and connect a .01 capacitor from the twisted leads to chassis ground. Connect the other leads to the antenna terminals of the television set. This shorted stub presents a high impedance across the antenna terminals at the television frequencies, so it doesn't degrade the performance of the set, but at the 15.75 kHz sweep-oscillator frequency it looks like a dead short to ground.

**fig. 2. Shorted 300-ohm stub across antenna terminals prevents radiation of tv sync pulses.**



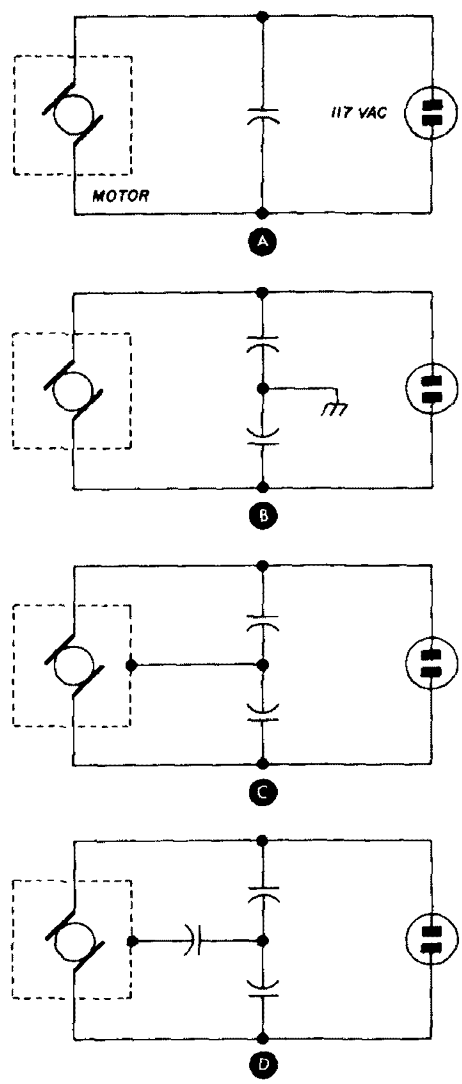




**fig. 3. Methods for installing filter capacitors across noisy electric motors. Capacitors may be from .002 to 2 μF, see text.**

Tv sync interference is sometimes radiated by the wires that run to the base of the picture tube. This can be cured by wrapping the wires with aluminum foil and grounding the foil to the tv chassis. (Don't wrap the high-voltage lead going to the side of the tube.)

Severe cases may require shielding the B+ wiring and additional bypassing in the sweep-oscillator stage, or lining the inside of the cabinet with screening, but at this point it would probably be worthwhile to consider a newer television set.

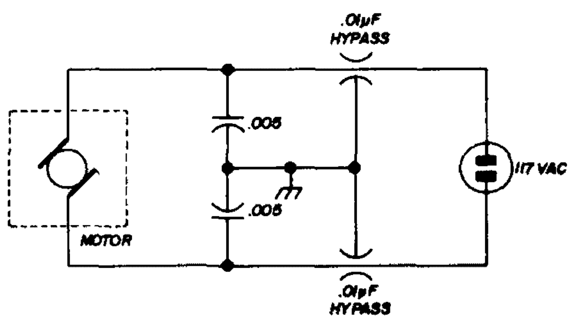



**fig. 4. When the simple filter circuits shown in fig. 3 don't suppress the noise sufficiently, .01 μF Sprague Hypass capacitors should be added in each side of the line.**

## electric-motor interference

Most of the newer motor-powered appliances have built-in noise suppression devices but many older units do not. Series or universal motors are usually used in small household appliances such as electric razors, vacuum cleaners and food mixers. Induction motors are used in larger equipment: refrigerators, washing machines, ventilating fans, etc. Induction motors with split-phase starting have no commutators and are not noise makers unless they are exceptionally dirty. However, many motors use a commutator for starting only, so radio interference is only caused for a few seconds.

To eliminate radio noise from a commutator-type motor, have the commutator turned down on a lathe, re-seat the brushes, install a well-grounded filter as close to the motor as possible, and keep the commutator scrupulously clean. However, it's usually difficult to keep commutators in series motors in good condition because of their small size, so some sparking will inevitably result.

The most effective radio-interference filter is a large capacitor installed as close as possible to the noise source. Capacitors may be used in any of the ways shown in fig. 3 – values will vary from 0.002 to 2 μF. For portable appliances the filter capacitors are limited in size by the requirement that possible current to ground through the capacitor may not exceed 0.3 mA. This protects the user from appreciable electrical shocks. The capacitors used in fig. 3C, for example, should not be larger than 0.005 μF: those used in fig. 3D should not exceed 0.01 μF.

In fixed installations which are permanently grounded the capacitors can be much larger, with a value of 2 μF as the practical upper limit. An excellent capacitor for this purpose is the 2-μF mylar capacitor sold by J. W. Miller. This capacitor, the Miller type 7804, is rated at 220 volts ac/dc and is designed to withstand transient surges up to 1000 volts.

If the simple capacitor filters shown in fig. 3 do not reduce radio interference to an acceptable level, add a 0.01 μF Sprague Hypass capacitor in each of the power leads as shown in fig. 4. These coaxially-constructed capacitors are effective radio-interference suppressors up to 150 MHz.

Most cases of motor-generated radio interference can be suppressed with the circuits of figs. 3 or 4 but occasionally you may run into a stubborn machine that requires the added suppression of an LC circuit. The best circuit for this purpose is the one recommended for your receiver (see fig. 1), although there are a number of compact commercial units available. A summary of commercial line filters and their wattage ratings is shown in table 1.

## fluorescent-light interference

Fluorescent lights, especially fluorescent desk lamps, are about the worst offenders found in any house. They are difficult to cure because they radiate radio interference through the ac power lines as well as through space. To solve ac line radiation build the circuit of fig. 5 into the base of the lamp and use an effective ground.

Direct radiation at distances closer than about 10 feet from the lamp is practically impossible to eliminate, although an rf shield of ¼" mesh screen may be placed across the opening of the reflector hood.

One type of fluorescent fixture that is particularly difficult to cure is the two-bulb lamp that uses pushbutton switches rather than a starter. One suppression method that is effective in these lamps is shown in fig. 6. Disregard B1, C1 and RFC1 if only one bulb is used: the start switch will be an spst type.

## other interference

Although electric motors, tv horizontal oscillators and fluorescent lights are the primary causes of rf interference, other devices can cause problems too – neon signs, diathermy, furnace ignitors and tv boosters – any device that is electric powered is a potential source of interference. Don't overlook that baby-bottle warmer, electric iron or electric blanket; they all contain a thermostat that generates a small arc each time it closes.

**table 1. Commercial interference filters for installation in the ac line.**

| wattage | V | manufacturer | type | model | application |
|---|---|---|---|---|---|
| 50 | 115 | Miller | LC | 7817 | electric shavers |
| 60 | 120 | Miller | LC | 7812 | teletype motors |
| 80 | 220 | Miller | L | 7878 | fluorescent lights |
| 220 | 115 | Miller | LC | 7818 | small appliances |
| 310 | 125 | CDE | LC | NF10280 | small appliances |
| 325 | 125 | Aerovox | LC | IN-106 | fluorescent lights |
| 550 | 115 | Miller | LC | 7815 | large appliances |
| 660 | 115 | Miller | C | 7816 | household appliances |
| 660 | 120 | Aerovox | LC | IN-42 | household appliances |
| 980 | 280 | CDE | LC | NF10431 | household appliances |
| 1100 | 220 | Miller | LC | 7880 | uncased, heavy duty |
| 1150 | 230 | Miller | LC | 7814 | business machines |
| 2200 | 220 | Miller | LC | 7881 | uncased, heavy duty |
| – | 120 | Aerovox | C | IN-105 | fluorescent lights |
| – | 120 | Aerovox | C | IN-27 | small appliances |
| – | 15 kV | Miller | L | 7875 | neon signs |

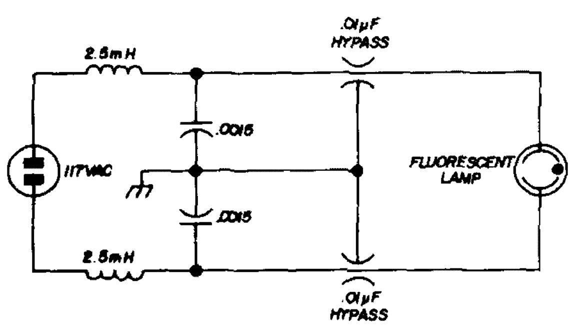



**fig. 5. Circuit for suppressing interference from fluorescent lamps.**

To eliminate interference from neon signs, replace any defective tubes, bond together all the conductive pieces in the field of the sign and check the insulation. High-tension filter chokes are available to eliminate neon-sign hash from the power line. One example is the J. W. Miller model 7875 insulated for 15,000 volts and designed for continuous operation up to 100 mA.

Thermostatic-arc interference is best suppressed with a filter capacitor as close to the contacts as possible. This type of interference is usually not too troublesome unless the contacts are especially dirty.

Rf heating units are another problem area. These units are found in diathermy machines and many industrial heating units. Since these units generate rf they must be licensed by the FCC, and must

meet FCC standards regarding spurious emissions. However, if the unit is tuned incorrectly, or operated improperly, severe rf interference can result. To cure interference, make sure the rf heating unit is grounded and shielded. Check frequency and harmonics; reduce drive to the final amplifier to reduce harmonic output, and if necessary, install traps and filters.

apply graphite-type belt dressing to the belt and bond the machines together and to ground.

One other type of interference that can give you fits is power-line noise – noise generated by arcing at dirty or broken insulators, tree limbs touching the wires or faulty transformers. However, unless you live near the ocean or in an area of heavy air pollution, the in-

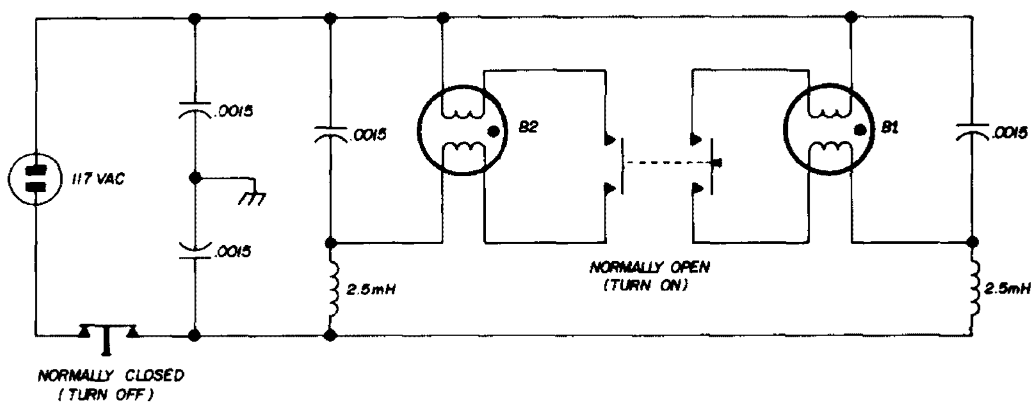



**fig. 6. Noise from fluorescent desk lamps is particularly hard to cure — this circuit usually does the job.**

Oscillating tv boosters are another source of interference, but not on the lower amateur bands. The first step here is to check the neutralization of the oscillating stage, check the dress on the input and output leads and provide more adequate shielding.

Remotely-controlled garage-door openers, if they use a superregenerative receiver, generate a lot of unnecessary rf interference that is best cured by replacing the culprit receiver with a non-radiating type.

Furnace ignitors, such as those used in oil burners, can generate an unbelieveable amount of hash that covers the lower amateur bands. Heavy-duty spark-plug suppressors and a line filter will usually quiet them down, but make sure that the burner unit, furnace and blower motor are all bonded together and grounded.

If your interference is emanating from a highly industrialized area, belt static may be part of the problem. To cure this,

sulators are usually cleaned by the rain. Crews from the power company try to keep branches trimmed back away from the lines, so that isn't often a problem, and modern oil-filled power transformers are designed for extremely long service. The result is that modern power distribution systems are *not* prominent noise makers, although they may act as a huge antenna for other noise sources. When they are found to be responsible for rf interference the power companies are usually helpful in finding and fixing the cause.

references

1. Frederick Littlefield, K1BQT, "Minituner SSB Receiver," *ham radio,* October, 1970, p. 16.

2. Doug DeMaw, W1CER, "A Noise-Locator Receiver," *QST,* June, 1966, p. 47.

3. W. R. Nelson, WA6FQG, "Electrical Interference," *QST,* April, 1966, and May, 1966.

4. Barron Kemp, "RF Interference Control Handbook," Howard W. Sams & Co., Indianapolis, 1962.

**ham radio**

# the rf bridge

Don Nelson, WB2EGZ, 9 Green Ridge Road, Ashland, New Jersey 08034

Sometimes an important idea goes unnoticed or is not sufficiently developed to gain wide acceptance. Such, I believe, is the case of the radio-frequency bridge. The rf bridge has been marketed for many years by the General Radio Company;[1] however, this precision instrument is probably too expensive for most amateurs. A moderately priced rf bridge, manufactured by Omega-t Systems,[2] has been available for several years. Oliver Swan, W6KZK, described the basic circuit of the rf bridge in an earlier issue of *ham radio.*[3]

Few hams seem to have recognized the advantages of the rf bridge over the simple vswr bridge. The rf bridge, for example, will allow you to optimize your antenna, thus reducing the dependency on a matching network. The rf bridge has other uses as well, some of which I'll discuss in the following paragraphs.

## the circuit

The instrument consists basically of a broadband noise generator coupled to a bridge network by a wideband 1:1 balun transformer. By carefully compensating for circuit strays, the bridge upper frequency limit can be extended to 450 MHz.

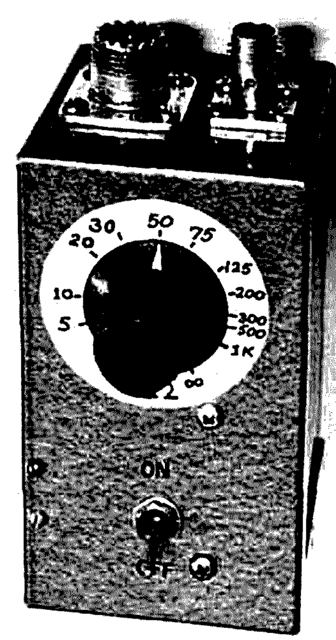
The circuit of fig. 1 was developed not without some difficulty, mainly in reducing circuit strays and constructing the balun transformer. In its present state of development, this circuit is useful to 220 MHz.

The noise generator uses a zener in an unstable (thus noisy) mode by operating it at low current. It will pay to experiment with the value of R1 for the highest noise level of your zener. When the noise-generator output is amplified by a two-stage broadband amplifier, the instrument is useful from about 1 to 450 MHz; again, the upper frequency limit is determined by how well wiring strays are compensated.

## construction

Simple construction was used, with parts mounted on a perforated board. Battery power was used for maximum utility. Wiring the bridge circuit is tricky, as might be expected with broadband equipment. If the layout shown is followed, you can expect good results. I feel there may be better layouts, and I'm sure

**Recommended replacement for the common vswr bridge — the radio-frequency bridge and noise generator.**

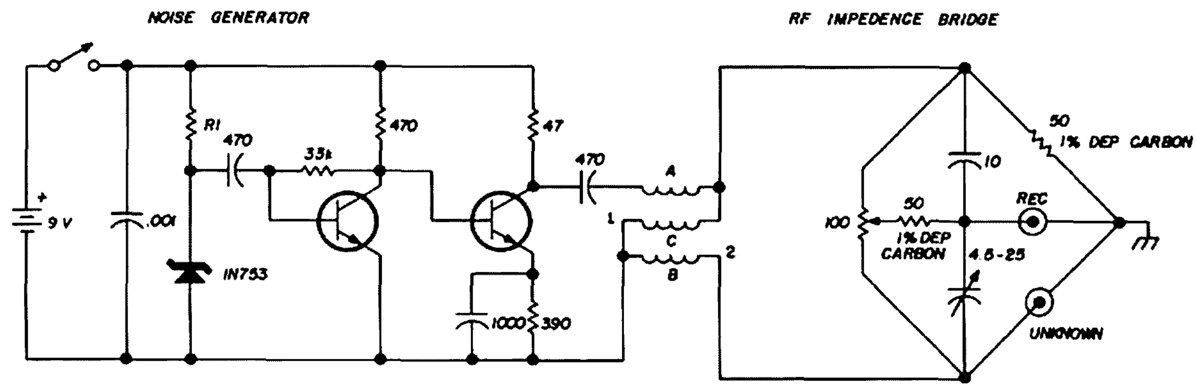


that every unit built will be slightly different with regard to compensation for circuit strays.

By far the most difficult part of the construction is the toroidal balun. The resultant transformer,[4] shown in fig. 1 has broadband characteristics that exceed those of the more common trifilar-wound units. Pay strict attention to details!

The bridge section was laid out with regard to uhf performance, keeping wires on one side of the bridge equal to those on the other. Wiring strays are compensated by balancing them with the exact capacitor combination that gives the best null. Because I have found the trimmer adjusts slightly differently on 6 meters and higher, I assume there are a few sneaky rf paths. One suspect component is the large carbon potentiometer. Our sophisticated doubts about the layout are unfounded below 30 MHz, however. (Solid relief for the unsophisticated worrier.)

This gem is self-contained in a Bud CU2103-A Minibox, ready to check antennas, receivers, quartz crystals, and other series-resonant circuits. You will, of course, need a receiver for null detecting at the frequency of interest.




NOTES:
1. R1 - CHOOSE SUITABLE VALUE FOR BEST NOISE OUTPUT (APPROX 2200 OHMS)
2. TRANSISTORS - 2N918, 2N3563, OR HEP 56
3. 0.375 OD TOROID FORM, INDIANA GENERAL CF 102, CORE MATERIAL 03

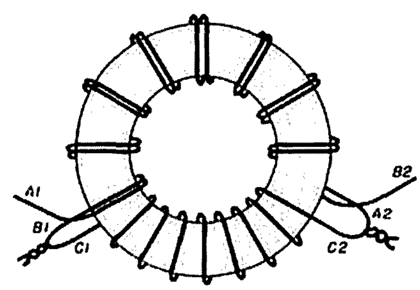


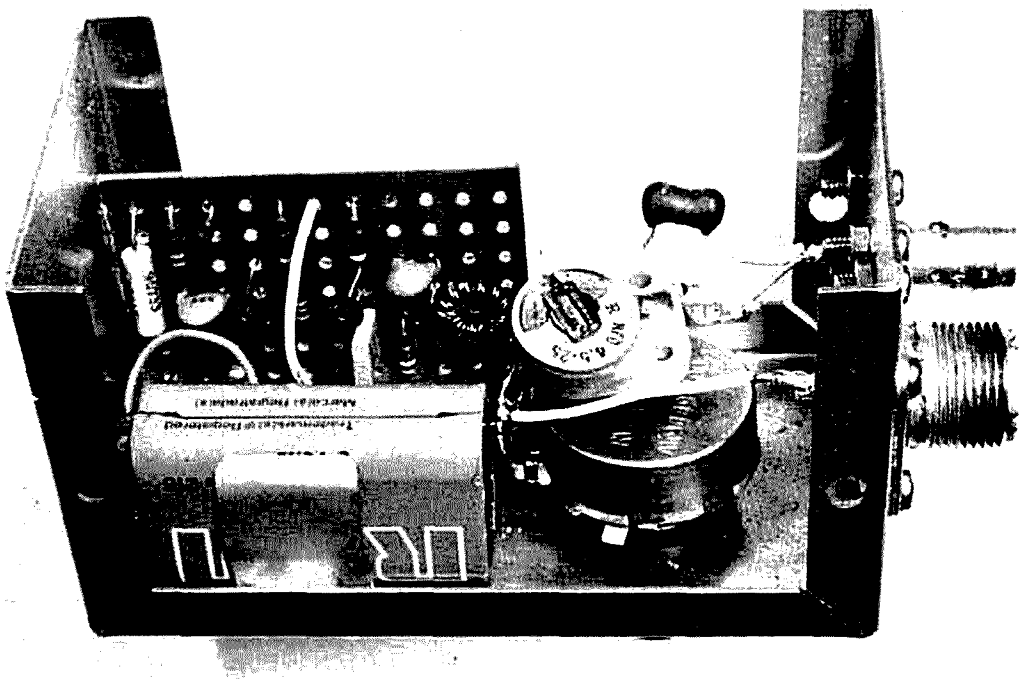

**fig. 1. Schematic of the rf bridge and noise generator, A. Windings A and B of the balun are no. 26 Formvar twisted 3 turns/inch before wrapping on core. Nine turns of the twisted pair are wound on the core. Winding C is also 9 turns of no. 26 Formvar, continuing the A and B winding direction and connecting A2 to B1.**

## calibration and use

In theory, if not in practice, the 100-ohm pot will balance any resistance placed in the "unknown" arm of the bridge. At one end of the scale is zero; at the other is infinity. Fifty ohms is midrotation with a linear pot. At 50 MHz and higher, I've found a rotational shift of the 50-ohm (rf) point. This means a special calibration check will be necessary at very high frequencies (vhf). Normally, for the hf range, the dial calibration will hold. The best null is at midrotational scale. Because the null deteriorates at the extremes of rotation, it is not worthwhile to use the instrument beyond a 20-to 300-ohm range.

Calibration is performed using noninductive resistors of known values placed, then nulled, across the UNK terminal, with a receiver connected to the

REC terminal. Carbon composition resistors are fine if values are known to 5%. Above 100 MHz, deposited carbon resistors are preferable because of their low inductance. The dial plate should be calibrated in the hf range, say 10 MHz. Trim the bridge capacitance for best null with a 50-ohm resistor and correct setting of the pot. Don't be too surprised if a 50-ohm resistor changes value through the vhf range.



**Parts layout, which should be followed closely for trouble-free results.**

## antenna matching

Tuning an antenna with a vswr bridge is a hit or miss proposition, because the vswr bridge confuses resistive and reactive impedances. I don't mean to imply that accurate tuning is impossible with the vswr bridge; but without a tedious procedure, the lowest vswr will probably occur at a frequency different than that of optimum transmission. The rf bridge technique eliminates the tuning error, and allows an accurate measurement of vswr once the antenna is correctly tuned.

1. First connect the rf bridge directly to the antenna or at an electrical half wavelength away from the antenna. An electrical half wavelength is different from the physical length of the wire. You can determine the electrical half wavelength with this bridge by setting the bridge to zero and placing a short across the end of the transmission line. Now cut small lengths from the line until a null is obtained at

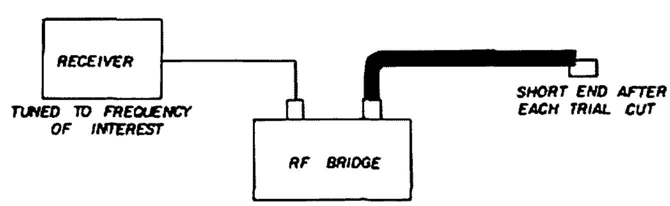



**fig. 2. Determining one-half wavelength of transmission line when using the rf bridge for antenna measurements.**

the frequency of interest (fig. 2). Using a half wavelength or multiple thereof effectively places the bridge at the antenna, thereby reducing transmission-line errors.

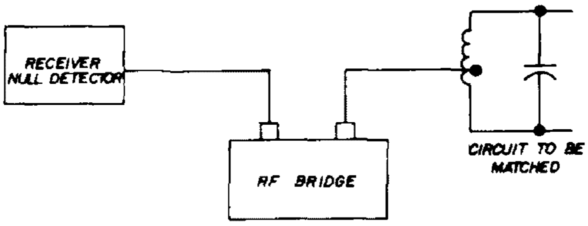



**fig. 3. Arrangement for matching input circuits.**

**2.** Tuning the antenna to a frequency is the next step. You will find its resonant frequency by a null on the receiver. A sharper null will be seen with the bridge adjusted to the impedance of the antenna system. Adjust antenna length until the null occurs at the desired frequency.

**3.** By adjusting the matching section, tune your antenna to the desired impedance as shown by the rf bridge.

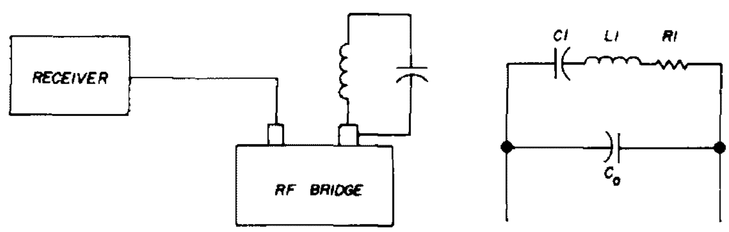
## other uses

Any series-resonant circuit can be checked with the rf bridge. This, you will recall, is the combination that cannot be dipped easily on a grid-dip oscillator. Place the LC combination across the UNK terminal with the bridge dial set to zero. Tune receiver for null. See **fig. 4.**

If a resistance is in series with L and C, the bridge will show its value. An interesting example of an R, L, C combination is the quartz crystal. While this bridge has limitations in crystal measurements, it is utilitarian. Set the dial to infinity (minimum noise for open circuit). Tune the receiver for an increase in noise at the resonant frequency of the crystal. Adjust the bridge for null. This value is the resistance of the crystal's RLC arm. In general, the lower this value, the higher will be the activity of the crystal.

The rf bridge takes over where the vswr bridge leaves off. To my embarrassment, the rf bridge singled out several mistakes in my station, as it may in yours. I feel certain that building this bridge will be the most rewarding project the experimenting amateur will undertake this year.

**fig. 4. Connections for checking series-resonant circuits. Network at right is the equivalent circuit of a quartz crystal.**




## receiver input matching

Provided you already have a receiver to act as a null detector, you will find the rf bridge invaluable for determining the optimum tap position for inputs to converters, preamplifiers, and receivers. The procedure is the same as before, except that the UNK terminal is now connected to a receiver input. With the bridge dial preset to the desired impedance, adjust the tap on the antenna coil for best null (**fig. 3**).

Grateful acknowledgement is made to Mike Ward, WB2YJK, for his efforts in the design of this project.

### references

1. General Radio Co., West Concord, Mass.
2. Omega-t Systems, Inc., 516 W. Belt Line Road, Richardson, Texas 75080.
3. Oliver Swan, W6KZK, "Impedance Bridge," *ham radio,* February, 1970, p. 67.
4. C. L. Ruthroff, "Some Broad-band Transformers," Proc. IRE, Vol. 47, August, 1959, pp. 1337-1342.

**ham radio**

# avalanche-transistor circuits

Common transistors operated in the avalanche mode make excellent pulse generators — here's a simplified theoretical discussion and some application ideas

E. Dusina, W4NVK, 571 Orange Avenue West, Melbourne, Florida 32901

**This article describes avalanche transistors in a simplified manner.** Basically, the avalanche circuit is a way to generate very high peak-power pulses from very small transistors. For example, a common 50-mW transistor can easily generate pulses with 100-W peak power. Such pulses not only have high power, but possess extremely fast rise times, which means they can generate harmonics into the lower microwave region.

Avalanche circuits are very simple and are useful in applications that require generation of harmonics, such as in frequency standards and in some types of low-level power generators at high multiplication ratios. When understood by the ham, more and more uses will be found for these circuits.

The "avalanche transistor" is a method of operating most any transistor rather than any special kind of transistor. This mode of operation is so simple as to be disbelieved until you build a circuit and see for yourself what magnificent pulses it will generate. **Fig. 1** is a schematic of a typical circuit. Most transistors will work in this circuit, and the rise time of the pulse is comparable to the highest frequency of oscillation for the transistor chosen. For example, a 2N918, 2N706, or the transistors in the μL914 integrated circuit generate harmonics to about 1000 MHz.

## how it works

The transistor in **fig. 1** operates similarly to a zener diode. For many small transistors, the breakdown (zener) voltage from the collector-to-emitter junction is about 70-90 volts. This voltage depends on the base connection, which can have

several configurations, ranging from $V_{ceo}$ (base floating) to $V_{cev}$ (base reverse biased). The collector-base junction breakdown voltage is shown in the device specification sheet as $V_{cbo}$. This para-

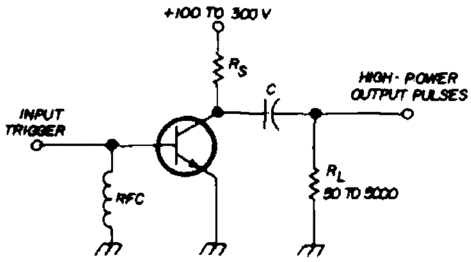


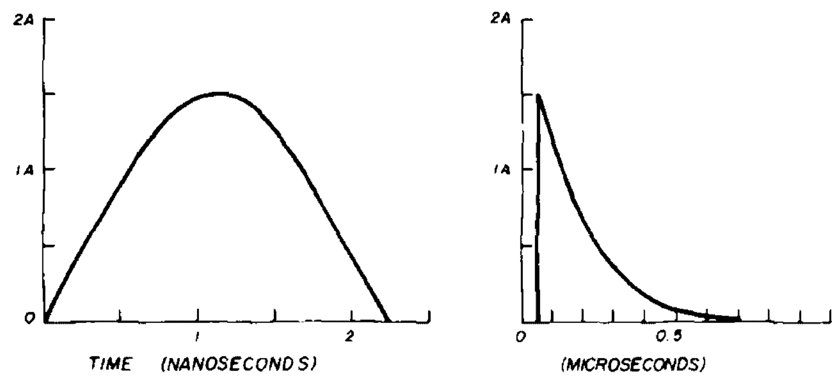

**fig. 1. Basic circuit of an ordinary transistor operating in the avalanche mode.**

meter specifies the breakdown voltage with the collector-base junction reverse-biased, and with emitter open-circuited. All of these breakdown-voltage parameters are given at a specific ambient temperature, usually 25° C.

When the transistor base is connected to the emitter, the transistor remains off, and only about 0.5 mA of collector current flows. This low current is due to the zener action and the current-limiting resistor, $R_S$.

When the base voltage increases to more than about 0.5 V above the emitter voltage, avalanche breakdown occurs. The transistor is then virtually a short circuit, and it will pass a high current. The voltage across the collector-base junction collapses in about 1 nsec for many vhf transistors.

**fig. 2. Approximate current pulses when load resistor $R_L$ = 50 ohms. Time constant, $R_LC$, is approximately 1 nsec in A; 100 nsec in B.**




## charging circuit

When avalanche breakdown occurs, the current flow due to the charged capacitor, C, is limited only by the load resistor, $R_L$, and the small internal resistance of the transistor. If C were very large, the transistor would literally explode. Capacitor C is small in value to keep the current flowing just a little longer than its rise time, since nothing is usually gained by keeping the current flowing longer.

Capacitor C charges to the zener voltage at which the transistor operates; and during discharge, the current pulse in $R_L$ is that voltage (70 V for example) divided by $R_L$, or about 1.4 A. The peak-pulse power is 70 V times 1.4 A, or about 100 W. The current flows for about two $R_LC$ time-constant periods. In this example, $R_L$ is 50 ohms and C is 20 pF; so the time constant is about 1 nsec. The pulse across $R_L$ will therefore look somewhat like that shown in **fig. 2.**

This powerful pulse is about equal to a single half wave of a 250-MHz 100-watt cw transmitter. Due to the high power, harmonics will be heard above 500 MHz.

Avalanche is controlled by a synchronizing signal introduced into the base-to-emitter circuit. Avalanche action occurs each time the input voltage rises above the 0.5-V trip point. Shown in **fig. 3** is a typical firing sequence when a sine-wave signal triggers the circuit.

If the input signal were 10 MHz, for example, this circuit using a 2N918 could generate harmonics to 1000 MHz. Zero

beating the 10-MHz source with WWV would yield a signal at 1000 MHz accurate to better than 200 Hz. Not bad for a one-transistor circuit!

## design details

Reviewing fig. **1**, let's look at the input circuit from base to emitter. A coil was used, and rf was impressed across the coil. A low-value resistor could have been used. So long as the total resistance in the base circuit is less than a few-hundred ohms, the circuit will remain off until fired by the input signal. Remember that the zener current comes out the base.

If 0.4 mA flows, for example, the total resistance in the base must be low enough so that 0.4 mA won't cause more than about 0.2-volt drop across the base-emitter junction. More resistance will bias the circuit so that erratic firing will occur from noise or stray pickup. Obviously if the base resistance (or impedance, if it's a coil) is too low, it will be difficult to drive a 0.5-volt signal across the base-emitter junction; so don't go overboard on the low side either.

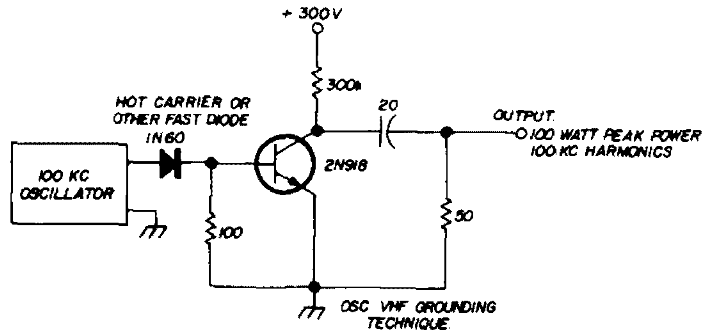



**fig. 4. Example of a diode used to protect drive circuit.**

### *erratic triggering*

If a tricky application causes erratic firing due to pulse pickup, which you can't suppress, a bias of 1.5 V from a battery in series with the base resistor will restore stability. The input trigger voltage must overcome the added bias in this case, and a 2-volt trigger will be needed; but immunity from false-firing signals will be greatly enhanced.

If you don't use good uhf grounding technique, the pulse can feed back into the drive circuits. For example, the pulse can easily cause a 10-20-volt drop across a 1-inch emitter-lead length. That voltage will couple back into the trigger circuit and possibly upset it. A shorter emitter lead is obviously the answer; but if the shortest lead you can get still fouls up, use a diode to block the spike, as is shown in fig. 4. The ground should be within about 1/8 inch of where the lead exits from the transistor can for best results.

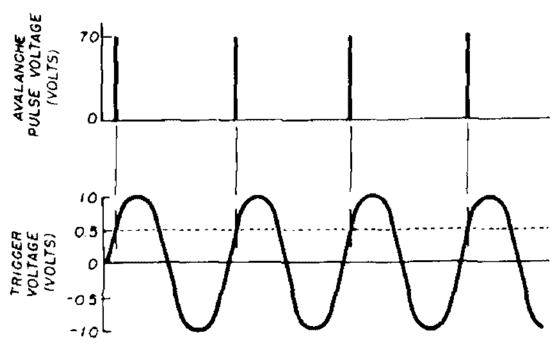



**fig. 3. Relationship between avalanche-pulse voltage across the load resistor and a sine-wave trigger voltage.**

## collector circuit

Now, let's look at the collector circuit to complete our design comments. The power dissipation of the circuit must be kept below the rating of the transistor. The zener voltage (70 V) times the current (0.4 mA) gives the dissipation (28 mW). The current in the transistor need

only be large enough to recharge capacitor C between firing times. The recharge time constant is $R_SC$.

Allow two time-constant periods between firings for peak output. The circuit will fire at much less, however; it will fire even if C doesn't fully recharge, but reduced pulse amplitude will result, and stability will be a bit more fussy.

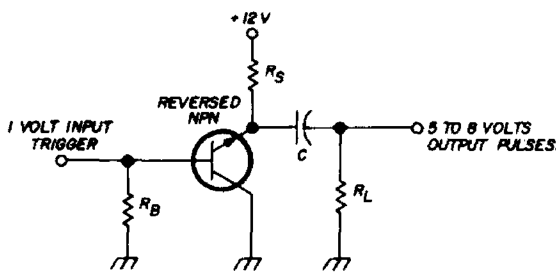



**fig. 5. Typical circuit to permit use of low-voltage power supply. Transistor emitter and collector are reversed for this application.**

## power dissipation

Resistor $R_S$ must be chosen to yield low current to stay within power-dissipation ratings as well as high firing by using a regulated voltage just larger than the zener level, and a low value for $R_S$ to give high recharge current but low idling current. Power loss during discharge plus idling dissipation must remain within transistor ratings.

If you make C too large the junction can melt during a pulse or after a few pulses. The maximum C value must be found experimentally. More than .001 $\mu$F gets risky, but up to .01 $\mu$F can be used with some transistors. Expect to melt a few semiconductors when playing with a large C value.

## low-voltage applications

A common problem with these circuits is the requirement for a high-voltage source. Very high pulse power is obtainable when high voltage is used, but many solid-state circuits use much less powerful pulses. By merely reversing the transistor (inverting it), we can get zener action at low voltage.

Fig. 5 shows a transistor with emitter and collector transposed. The zener voltage will be about 5-10 V, and the circuit may be operated from a 12-V dc supply.

## avalanche oscillator

If resistor $R_S$ is made adjustable, the circuit in **fig. 6** can be made to fire at a submultiple of the input signal, and you'll have what is known as an avalanche oscillator. Synchronization is a bit tricky, especially if the 12-volt supply isn't well regulated, but this one-transistor circuit can divide just like flip-flops or neon-bulb RC oscillators.

Because of their narrow width, the pulses generated by these circuits are hard to see on regular oscilloscopes, but usually something will show up on a 4-MHz scope in a darkened room.

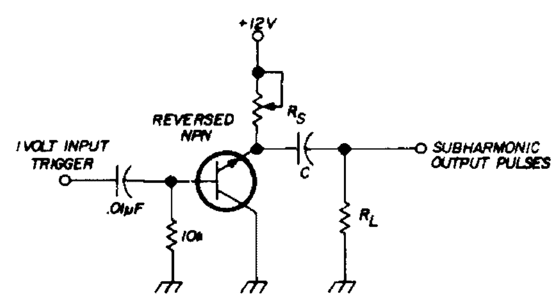



**fig. 6. A free-running avalanche oscillator synchronized to a subharmonic of the input signal.**

## conclusion

Well, there you have it – brief, but with all the basics. These little gems will allow you to generate everything from a controlled, hefty, 1000-MHz calibration signal to one of the dangdest noise makers going.

Considerable circuitry can be eliminated in some applications. An avalanche circuit on the output of my 5-kHz calibrator produces S-9 signals up to 10 meters. This makes it mighty easy to find my own signal when calibrating.

**ham radio**

# low-power transmitter and indicating wavemeter

These ultrasimple circuits will introduce you to micropower communications — and a lot of operating fun

Alf Wilson, W6NIF, 3928 Alameda Drive, San Diego, California 92103

An interesting and challenging part of amateur radio that deserves more attention is communications using very low-power equipment. This is known as "QRP" operating. The name is derived from the international list of radio Q signals and means, literally, "reduce your power."

A most informative article on this operating mode is by Art Child, W6TYP (reference 1). Among other accomplishments using very low power, Art has made a 354-mile, two-way radio contact using a power input of 354 microwatts. This figures out to be *one-million* miles per watt. Undoubtedly this record has since been shattered, but it typifies one of the objectives of the QRP Club, which has several thousand members throughout the world.*

One of the problems when working with extremely low power is adjusting the transmitter for maximum output. With power of the order of microwatts, a pick-up loop and flashlight-bulb arrangement just doesn't have sufficient sensitivity to give a good indication of maximum transmitter performance. A receiver with an S-meter could be used for peaking the output of a micropower transmitter, but many hams don't have receivers with this refinement.

In QRP work, every bit of available rf energy is important. The slightest misadjustment of the transmitter can mean the difference between success and failure in making a contact. This article describes a simple indicating device that will aid the QRP enthusiast in adjusting the transmitter for maximum output. Also included is a schematic of a typical QRP transmitter for those who might wish to try their luck in the gnat-power regime.

*The QRP Club recording secretary is F. Behrman, K7LNS, 3425 S. E. King Road, Milwaukie, Oregon.

## field-strength indicator

The circuit of fig. 1 is simplicity itself. The arrangement to the left of the transistor is nothing more than a wideband diode detector and an rf-pick-up wire. This circuit has appeared in various forms in many amateur publications. The transistor circuit improves things by allowing a milliammeter, rather than a more expensive microammeter, to be used as an indicating device. The transistor operates as a current amplifier. With an input of 100 microamps applied to the emitter, a range of 0-100 microamps (approximately) can be measured with a 0-1 dc milliammeter. The circuit could be made more sophisticated to improve measurement accuracy, but all we're interested in is adjusting a QRP transmitter for full output power. Thus a relative indication is adequate for this purpose.

## construction

I built my QRP field-strength indicator on a 2-inch-square piece of perf board, which fits nicely inside a universal meter case. I used an aluminum case, which costs a little more than the steel variety, but the aluminum is nonmagnetic and much easier to work with.* The finished unit has many other uses around a ham station. It's a great help when neutralizing rf amplifiers, or for making antenna adjustments. It can also be used as a detector for a-m phone signals.

A short length of vertical wire as a pickup antenna is adequate for use with transmitters of a few milliwatts or so. Sensitivity can be improved by using a 2-turn loop about 2 inches in diameter in place of the vertical antenna. Each pickup antenna (rod and loop) may be soldered to a phono plug for easy substitution. The loop works best when using the unit for neutralization.

Thus the indicating device is really nothing more than an improved version of the loop and flashlight bulb we used for tuning up our tnt oscillators in the olden days. Those who have tried to peak the power output of a QRP rig while trying to discern the barely perceptible glow of a flashlight bulb in a darkened room will appreciate this little gadget.

*Available from Allied Electronics, 100 N. Western Ave., Chicago, Illinois 60680. Stock no. 42F8542; $1.80 each.

## qrp transmitter

Most serious QRP hams try to build the simplest possible transmitter using a bare minimum of components. Although the oscillator shown in fig. 2 could be simpler, I've found this circuit to be the best compromise between reliability and simplicity. For example, many QRP rigs feed the antenna through a fixed capacitor tapped directly onto the tank coil, thus eliminating the coupling link and its mounting hardware. The direct-feed method, while certainly not the best example of good engineering practice, is okay for some situations. However, it's probably one of the most efficient harmonic radiators one could come up with. It seems improbable that a transmitter running such low power could cause interference to other services, but it can and has happened. In my case this little rig, running about 50 mW input on 21 MHz and using the direct antenna-coupling method, put a healthy 5th harmonic signal into the middle of the fm broadcast band on a neighbor's receiver. The problem was remedied by changing to the inductive-coupling method shown in the drawing.

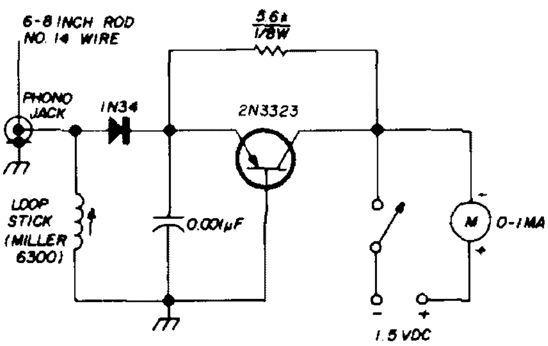



fig. 1. Rf indicating device for QRP transmitter tuneup. Current amplifier increases sensitivity of the milliammeter by a factor of ten.

## construction

This little oscillator was built on a piece of perf board, which was trimmed to press fit inside a small metal box such as throat lozenges come in. The box

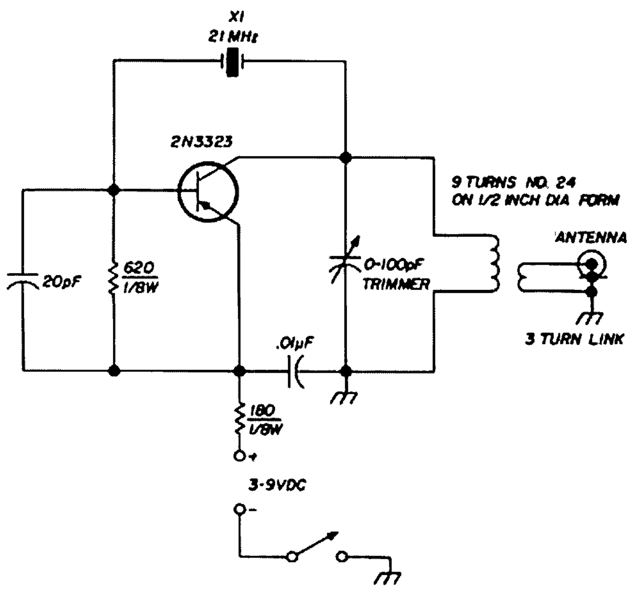



**fig. 2. Typical QRP transmitter for 15 meters. It might be necessary to experiment with the feedback component values to obtain oscillation consistent with transmitter loading and good keying.**

measures 1 x 2 x 3½ inches and has a hinged lid. I mounted the crystal holder, tuning capacitor, and antenna output jack on the lid. When closed, the box makes a tidy package that fits in a shirt pocket. A coat of aluminum spray lacquer completes the job.

With small-signal transistors now available by the sackful for very little money, you could probably substitute a less-expensive device for the Motorola 2N3323s used in these circuits. However, at 60 cents each I felt that the 2N3323 was more than worth the price. It exhibits good power gain well into the vhf region and works well in both applications shown.

## final comments

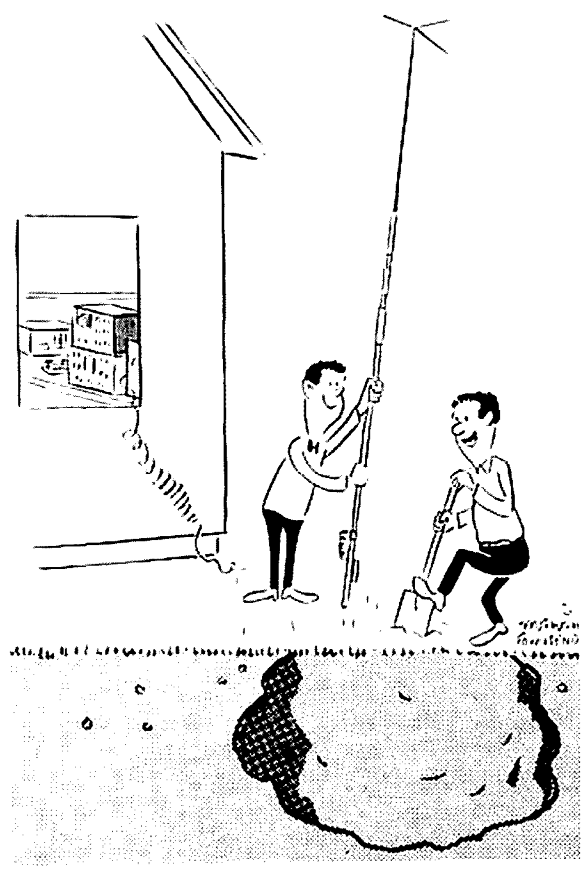
It seems that there are two camps among QRP operators, the purists and those not so pure (I happen to be one of the latter). The purists strive to limit their equipment to the barest of essentials, including receiver and antenna. While this is a very noble attitude, it requires the patience of a monk and unlimited time for operating. I tried the purist approach for awhile and managed to work one station. This fellow was operating mobile and heard my minibeeper when driving past the house. The QRP rig was loading a bedspring for an antenna. The contact lasted all of three blocks, so I finally gave up and connected the rig to the regular station antenna.

Working stations in the QRP mode under austere conditions is certainly an admirable achievement. If that's your bag — fine. But I've found that a good antenna and selective receiver make QRP operating less nerve wracking and much more fun.

### reference

1. Arthur Child, W6TYP, "QRP — A New World to Conquer," *73*, May, 1969, p. 46.

ham radio



**"This looks like an ideal spot."**

# a synchronous-phase afsk oscillator for RTTY

This circuit meets all requirements for generating the high-quality tone sinusoids necessary when using ssb transceivers in the RTTY mode

D. H. Phillips, W6FOO, 215 Carlisle NE, Albuquerque, New Mexico 87106

**Audio frequency shift oscillators** used to encode RTTY signals must operate in conformance with good design practice to avoid adjacent-channel interference. When an afsk oscillator is used with an ssb exciter to transmit carrier-shift fsk on the hf bands, the afsko must generate stable, sinusoidal tones. The ratio of space-tone to mark-tone amplitude must be variable to allow compensation for the combined frequency response of the microphone audio amplifiers and the crystal filter. This capability is required for adjusting the afsko so that the carrier power will be constant for both space and mark transmission. Finally, the most important afsko requirement for minimizing spurious signal generation on hf RTTY is that the transition from one tone to the other be accomplished without a phase discontinuity. The synchronous-phase afsko described here has all the required capabilities for generating the high-quality afsk waveforms required for transmitting carrier-shift fsk using an ssb exciter or transceiver.

The oscillogram of fig. 1 shows the input-output characteristics of the synchronous-phase afsk oscillator. A positive input voltage greater than 0.7 volt constitutes a space command. The

mark tone is generated for an input voltage less than 0.7 volt.

## control circuit logic

The afsko uses a Signetics SP620A flip-flop IC.* The 620A is a dc-triggered, master-slave flip-flop that can be switched synchronously by using the J and K inputs together with a clock. The rising clock pulse cuts the slave off from the master. As the clock pulse rises still higher, it allows the logic at the J and K inputs to be set into the master. Then when the clock pulse returns to its low level, the state of the master is transferred to that of the slave which, in turn, sets the output levels. Transfer gate and master flip-flop gate thresholds are separated by sufficient voltage to guarantee that input and transfer cannot occur simultaneously.

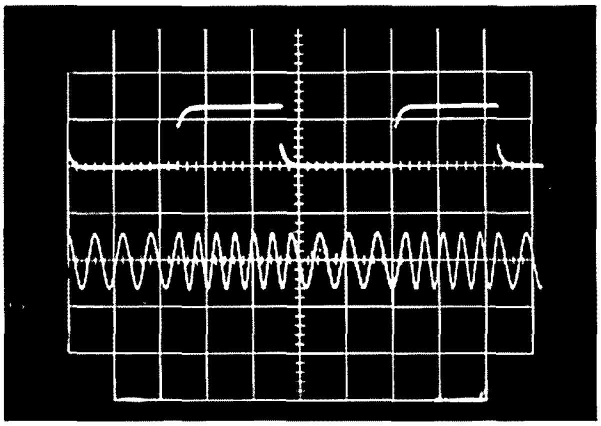


**fig. 1. Input-output characteristics of the synchronous-phase afsk oscillator. Top is the input signal, 5 volts/division. Bottom is output signal 1 volt/divison, 1650 Hz mark, 2500 Hz space. Horizontal scale is 1 millisecond/division.**

The flip-flop logic diagram, composed of AND and NOR gates, is shown in fig. 2. Fig. 3 shows the input, clock, and output waveforms for synchronous operation. The inputs can be switched at any time, but the outputs switch to obey the input command *only* on the trailing edge of a positive clock pulse. The flip-flop functions as an electronic delay line. When a change in the input signal occurs,

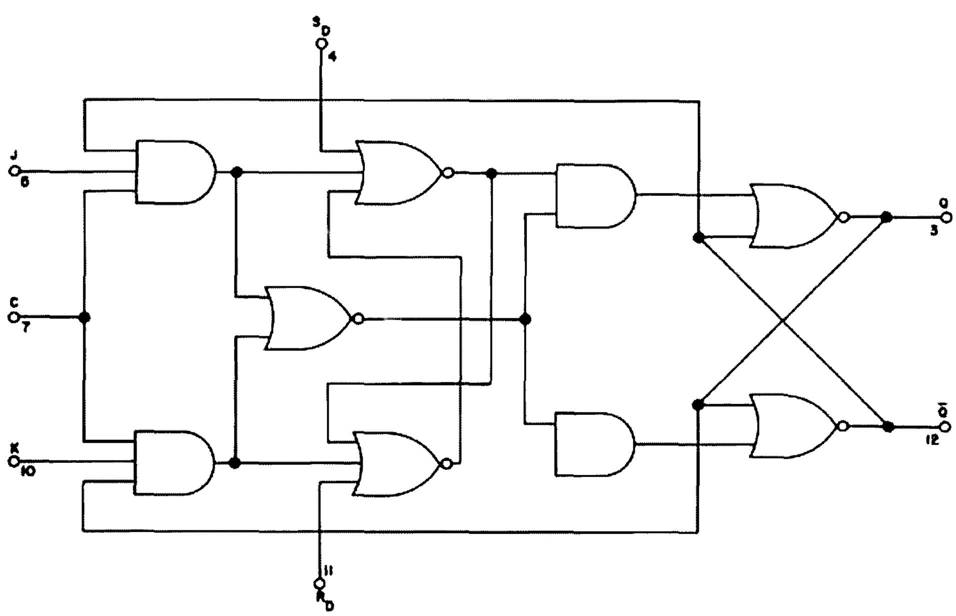



**fig. 2. Logic diagram for the Signetics SP620A flip-flop IC used in the afsko control circuit.**

*The SP620A (price $3.50) is available from Compar Corporation in Huntsville, Alabama; Glendale, California; Hamden, Connecticut; Seattle, Washington; and other locations.

the new signal is delayed by the flip-flop until the next clock pulse trailing edge, at which time the new signal appears at the Q and $\bar{Q}$ output terminals. The delay time

is variable and can range from 65 nsec to infinity, depending upon the clock pulse-to-pulse interval.

## the afsko circuit

Figs. 4 and 5 are the functional block diagram and schematic of the synchronous-phase afsko. The input signal to the afsko is processed by Q5-Q6, which provide complementary inputs to the flip-flop. Q5 is a threshold detector that conducts for any input voltage greater than 0.7 volt. Q5 is off for any input voltage less than the 0.7-volt threshold. D8 protects Q5 by preventing excessive reverse bias being applied to the Q5 emitter-base junction. D8 also allows the afsko to be driven by a source having an ac-coupled output, since D8 provides a discharge path for the driving-source output capacitor. S1 is a reversing switch that allows the space and mark signals to be inverted; thus the afsko can be used on either lower or upper sideband.

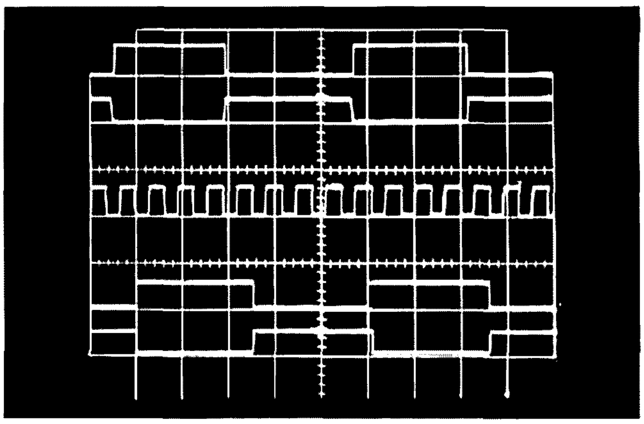


**fig. 3. Switching characteristics of the SP620A IC. Vertical scale, all traces, 0 to + 6 volts, 10 volts/division. Horizontal scale is 1 millisecond/division.**

## sine-wave oscillator and buffer

The basic sine-wave oscillator[1] consists of the complementary amplifier (Q3 and Q11) and a tuned LC circuit. The tuned circuit, composed of L1 and one of three capacitors ($C_{sn}$, $C_{sw}$, or $C_m$), is switched in parallel with L1. Only one of the three capacitors is in parallel with L1 at any given time. Q1 switches the mark-tone capacitor, $C_m$, into and out of the oscillator circuit. S2B selects the desired space-tone capacitor; $C_{sn}$ for narrow-shift

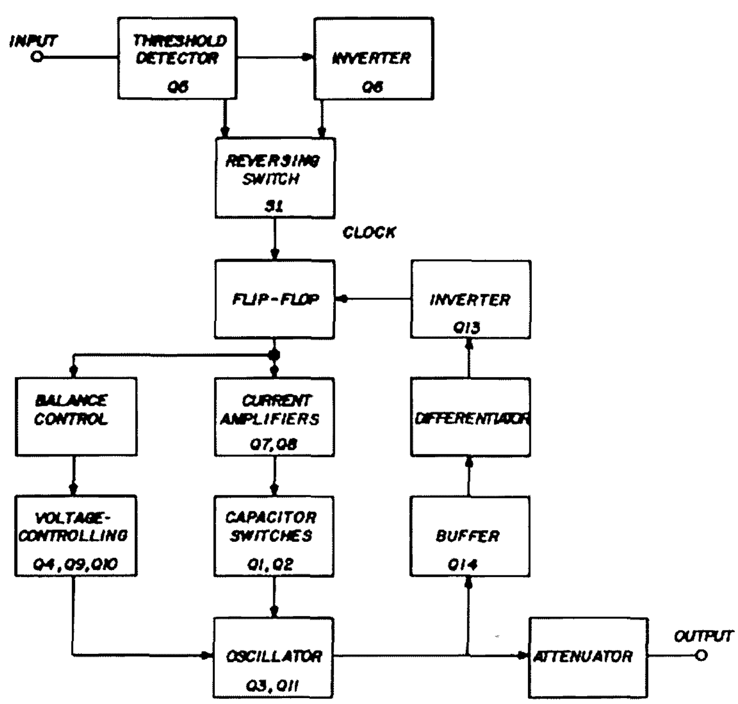
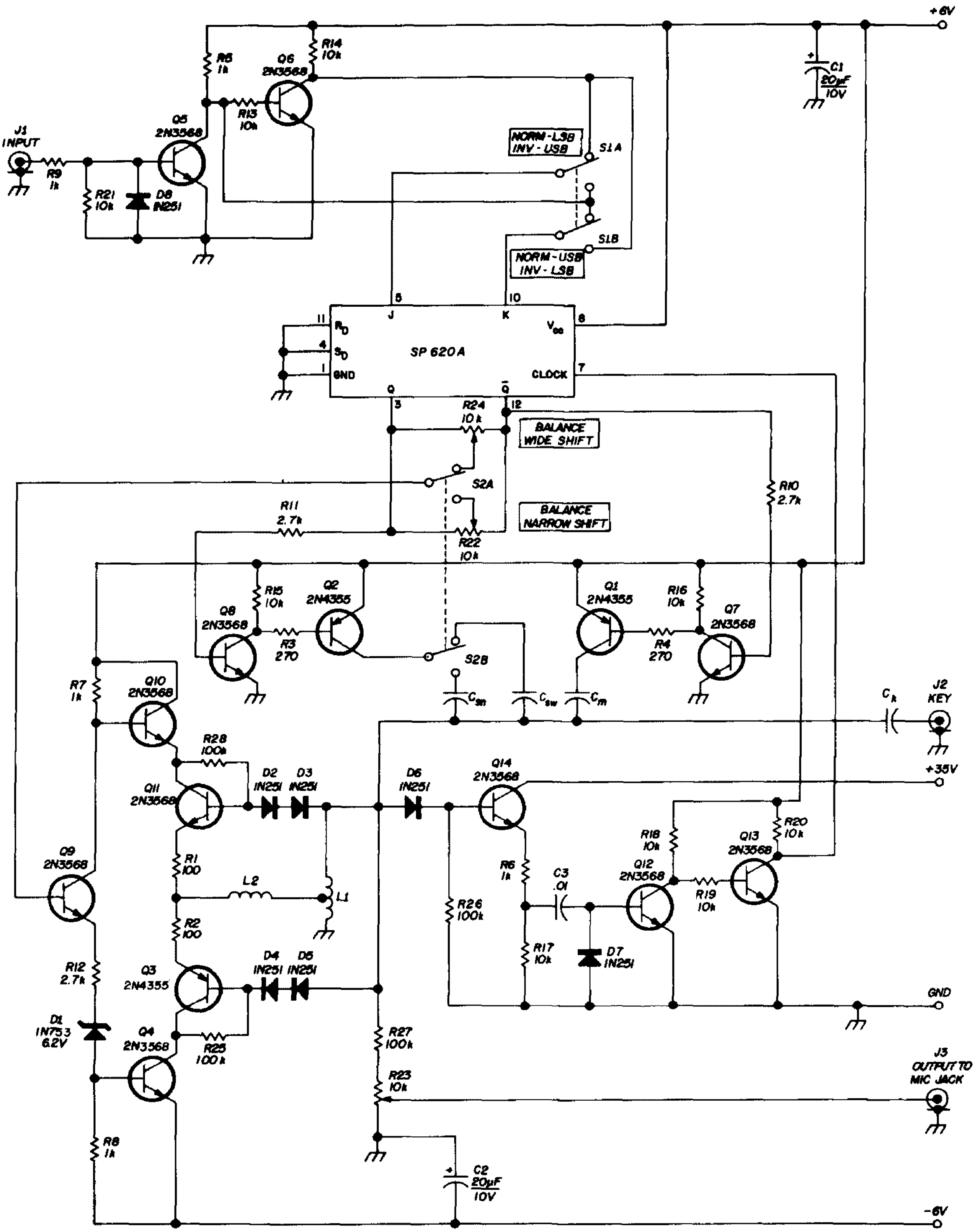



**fig. 4. Block diagram of the synchronous-phase afsko.**

**fig. 5. (right) Synchronous-phase afsko schematic. Capacitors Csn, Csw, Cm must be selected for space and mark tone frequencies. See text.**




$C_{sn}$, $C_{sw}$, $C_m$ select for desired narrow-shift space, wide-shift space, and mark tone frequencies. See fig. 11 and table 1 for capacitance values. Use capacitors having a voltage rating of $>$20 v

$C_k$ select for the desired cw identification frequency shift

$J_1$-$J_3$ type C-11, ¼-in. phone-jack, or equivalent

$L_1$ 88 mH

$L_2$ 22 mH. Use only one winding of an 88-mH toroid

$D_1$ 1N753, 6.2 V zener (price $1.05) or equivalent

$D_2$ - $D_8$ 1N251 (30 Volts PIV. Price 32¢) or equivalent

$Q_1$ - $Q_3$ 2N4355 (price 55¢) or equivalent

$Q_4$ - $Q_{14}$ 2N3568 (price 32¢) or equivalent

SP620A Signetics flip-flop (see Text)

operation, or $C_{SW}$ for wide-shift operation. Q2 switches the selected space-tone capacitor in parallel with L1 only during space intervals. Actually, these capacitors are switched to the +6 volt power supply, but C1 provides an ac short circuit from the power supply to the grounded side of L1.

The oscillator drives a buffer stage, Q14, which prevents oscillator loading.

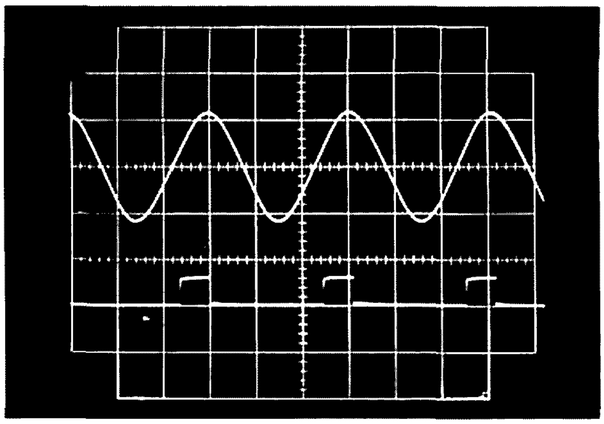
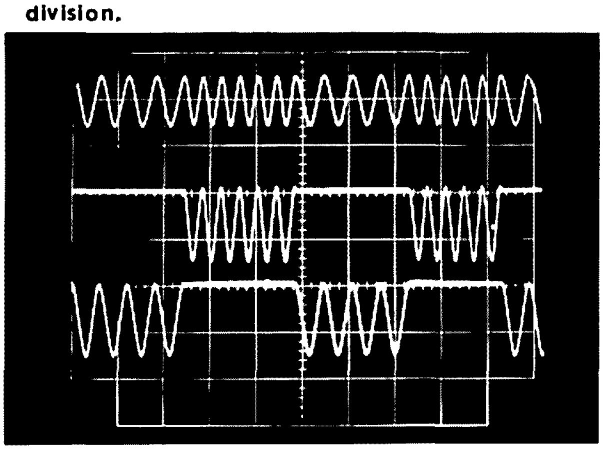


**fig. 6. Time relationship between clock pulse and afsk oscillator output-signal waveform. Clock trailing edge occurs at positive peak of output waveform. Top is output signal, 0.5 volt/division. Bottom is clock signal, 10 volts/division. Horizontal scale is 0.2 milliseconds/division.**

The buffer drives a differentiator, Q12. The function of this differentiating circuit is to generate a signal when the output sine wave reaches its positive

**fig. 7. Time relationship between output signal and collector voltage on mark-space capacitor switches, Q1 and Q2. Top is output signal, 1 volt/division. Center is Q2 collector, 10 volts/division. Bottom is Q1 collector, 10 volts/division. Horizontal scale is 1 millisecond/division.**



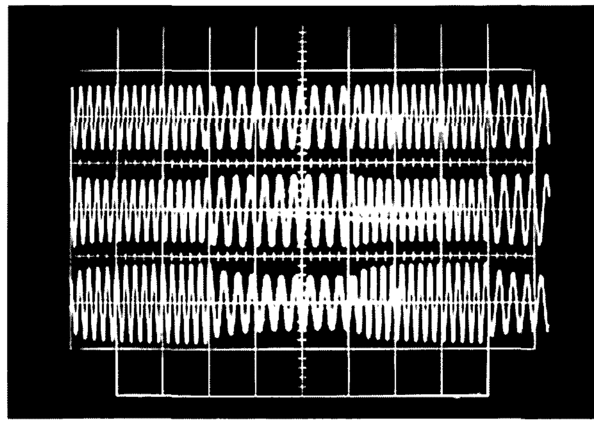


**fig. 8. Effect of balance control setting afsk oscillator output waveform. Mark, space, and switching frequencies are respectively 1650, 2500, and 100 Hz. Top is output waveform of properly balanced afsk oscillator. Center is balance control set fully counterclockwise. Bottom is balance control set fully clockwise. Vertical scale is 0.5 volts/division; horizontal scale is 2 milliseconds/divison.**

peak. At this point in time, Q12 is turned off and the inverter, Q13, is turned on, thereby providing a +6-to-OV transition signal to the clock terminal of the flip-flop. This high-to-low transition is the trailing edge of the positive clock pulse.

## timing

Fig. 6 shows the time relationship between the output sine wave and the clock pulse. The afsko is allowed to switch frequencies, from mark to space or vice versa, only at the time when the output sine wave reaches its peak; and only if the afsko input signal was changed during the previous period of the output waveform.

The Q and $\bar{Q}$ flip-flop outputs drive the current amplifiers Q8 and Q7. These current amplifiers provide the high base current required by the capacitor switches, Q1 and Q2. These switching transistors must support the circulating L1-C tank circuit currents.

At any given time, one of the capacitor-switching transistors (Q1 or Q2) will be on, while the other is off. Because the capacitor switching occurs at the positive peaks of the periodic output waveform, the collector-to-emitter voltage on Q1

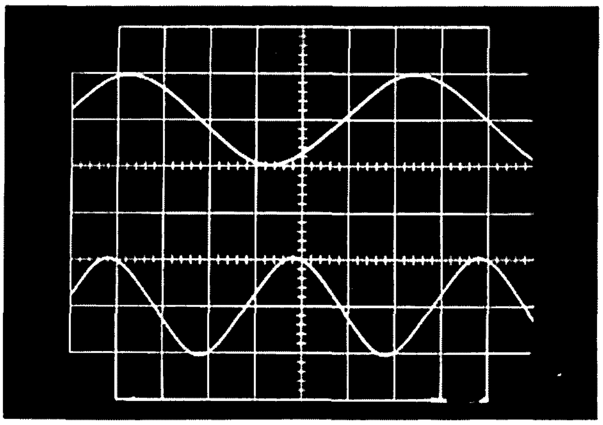


fig. 9. Output waveforms of the afsk oscillator showing quality of mark and space tones. Top is mark tone, 1650 Hz. Bottom is space tone, 2500 Hz. Vertical scale is 0.5 volt/division; horizontal scale is 0.1 milliseconds/division.

and Q2 is always zero or some negative value. Fig. 7 shows how the collector voltages of both Q1 and Q2 vary with respect to time as the afsko is electrically commanded to switch from space to mark and vice versa.

## narrow and wide shift

Switch S2A selects one of two balance potentiometers; R22 for narrow shift and R24 for wide-shift operation. The selected balance potentiometer provides a voltage to the voltage-controlling circuit composed of Q4, Q9, and Q10. This circuit controls the voltage that provides power to the sine-wave oscillator, so that the mark-to-space tone amplitude ratio can be varied. When the afsko is used to drive an ssb exciter, this ratio should be adjusted for equal space and mark carrier power. Fig. 8 shows the high-quality afsk modulation waveform generated when the balance control is adjusted properly. Also shown are conditions of unbalance that can be used to compensate for a nonuniform combined frequency response of the microphone amplifiers and the sideband filter. When unbalanced, the output waveform reaches its equilibrium amplitude within 2 ms (less than 10% of the bit time for 60-wpm teletype) after a change in tone frequency. R23 controls the afsko output amplitude, which can be varied from zero to 1 volt p-p. The output waveform is a high-quality sinusoid.

Fig. 9 shows the waveform quality for a mark tone of 1650 Hz and a space tone

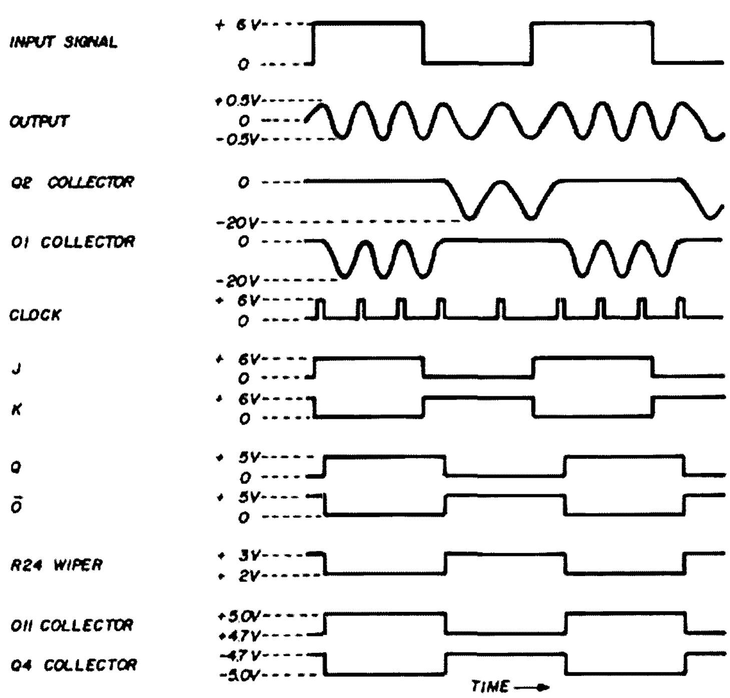



fig. 10. Timing diagram of the afsk oscillator.

**table 1. Recommended tone frequencies and capacitor values.***

| | upper sideband | | lower sideband |
|---|---|---|---|
| mark tone | 2500 Hz | mark tone | 1650 Hz |
| key-down tone | 2000 Hz | key-down tone | 1450 Hz |
| space tones | 2330 Hz (narrow shift) | space tones | 1820 Hz (narrow shift) |
| | 1650 Hz (wide shift) | | 2500 Hz (wide shift) |
| | $C_m$ = 0.045 μF 20V | | $C_m$ = 0.11 μF 20V |
| | $C_k$ = 0.036 μF 20V | | $C_k$ = 0.036 μF 20V |
| | $C_{sn}$ = 0.055 μF 20V | | $C_{sn}$ = 0.094 μF 20V |
| | $C_{sw}$= 0.11 μF 20V | | $C_{sw}$= 0.045 μF 20V |

*Note: Most capacitors have 20% tolerance, requiring hand selection for best frequency accuracy.

of 2500 Hz. The results of distortion-analyzer measurements made on the output signal from the afsko show a waveform distortion of less than 4% for any tone frequency below 3.5 kHz. The frequency stability is better than 0.2%, including the effects of balance control setting.

The waveforms at various points in the afsko when S1 is in the "normal (lsb)" position and S2 is set for wide-shift operation are shown in **fig. 10.**

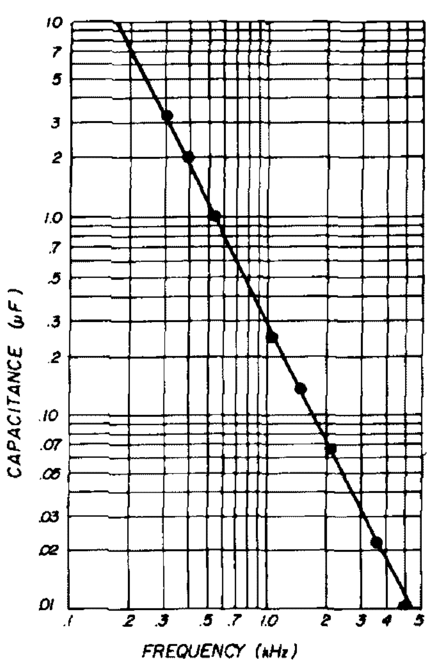



**fig. 11. Curve for determining capacitance for the three tone frequencies.**

## conclusion

The afsko can be operated with either lsb or usb ssb transceivers. The selection of tone frequencies should be made so that, in normal operation, the mark frequency doesn't change when switching from wide to narrow-shift. The three tone frequencies can be independently controlled by selecting the proper values for $C_{sn}$, $C_{sw}$, and $C_m$. **Fig. 11** can be used to select the proper capacitance value for the three selected tone frequencies. **Table 1** gives recommended capacitance values for usb and lsb operation. The recommended capacitor values aren't standard, so parallel combinations of two or more capacitors may be used, including very small capacitors for fine frequency adjustments.

The choice of semiconductors isn't critical, except that *silicon* transistors and diodes must be used. Bias conditions are designed to make use of the forward voltage drop (0.7 V) across a silicon diode, and capacitor switching must be done with low-leakage silicon transistors. The physical layout of the components isn't critical, and special construction techniques aren't required.

**reference**

1. D. H. Phillips, "An Audio Sinusoid Generator," *73,* September, 1969, pp. 88-89 and October, 1969, p. 131.

**ham radio**

# identifying unknown transistors

Simple methods you can use to classify bargain devices obtained from surplus sources

George R. Allen, W2FPP, 4059 Bay Park Drive, Liverpool, New York 13088

**Surplus transistors are readily** and inexpensively available from many sources. However, a large number of these transistors are either unmarked or marked with some undecipherable number. This is especially true of transistors on surplus computer boards.

An attempt to obtain data on these transistors by writing to manufacturers will result in a great deal of frustration, as manufacturers will not usually release specifications on specially marked devices. So if the transistors are to be of any value, the only solution is to identify and grade them. This article describes several such techniques that you can apply in your own workshop. Using simple circuits, you will be able to identify polarity (pnp or npn) and grade the devices into three general categories:

1. Audio (low beta).
2. Audio (high beta).
3. Rf types.

The techniques described here apply only to small-signal transistors and won't work for large-signal or power devices. The culls from the grading process will, in most cases, be of the latter type. These can be saved and used with other techniques for grading.

## circuit applications

For most practical purposes, transistors of the same polarity within a given

group (small-signal devices) can be used interchangeably. These general categories provide a starting point when building transistor circuits. The desired circuit is built, and the category of the required transistor is determined. Graded transistors of the proper category are then substituted into the circuit until optimum performance is obtained. In most cases, you'll find that just about all transistors in a given category will work equally well. The exception will be in circuits considered by the designer to the critical.

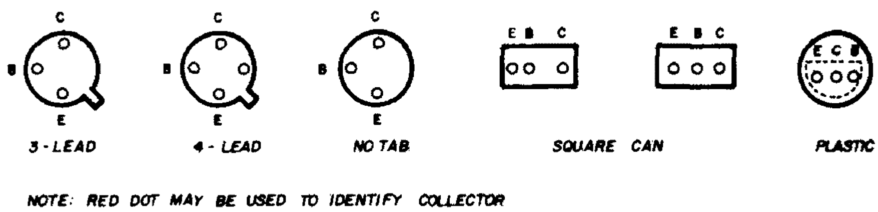



**fig. 1. Base connections for many small-signal transistors. Metal-case and plastic-encapsulated-device bases are shown.**

## pnp or npn?

Transistor polarity must be determined before the devices can be graded into their general categories. To separate them into pnp or npn types, use an ohmeter and refer to the base diagrams in fig. 1. Place the ohmeter negative lead on the base lead of the transistor to be tested (see **fig. 2**).

When the positive lead of the ohmeter is placed on either collector or emitter, the ohmeter should show a low resistance if the transistor is pnp. Now reverse the ohmeter leads; a high resistance should be indicated between base and collector and between base and emitter.

With the ohmeter negative lead on the base, a high resistance should be found between base and both emitter and collector if the transistor is npn. Reversing the leads should produce a low-resistance reading.

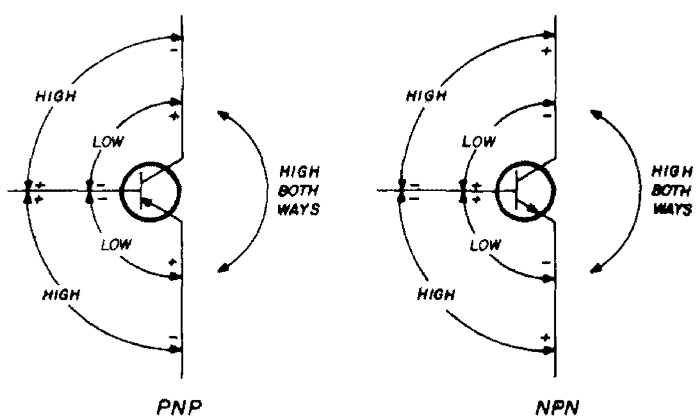



**fig. 2. Ohmeter connections for finding transistor polarity. Resistance readings determine whether unknown device is pnp or npn.**

It's always wise to reverse the ohmeter leads during this test to ensure the transistor isn't shorted. If a low resistance is measured between base and either collector or emitter, with the ohmeter in both normal and reversed positions, the transistor is shorted.

If, when testing a transistor, a low-resistance is found between base and one

lead; and a high resistance is found between the base and the other lead, the transistor is bad or you don't have the base identified correctly.

## determining lead connections

In cases where you're uncertain of lead connections, the following procedure may be used to determine the proper connections.

Looking at fig. 3, label the three unknown leads so you won't become confused during the test. Since you have three leads, there is a possibility of three arbitrary pairs. In the example shown, the pairs would be xy, xz, and zy. Select a pair at random, then connect the ohmeter so that the ohmeter reading is lowest. Now, short the third lead to each of the other two leads. If the resistance increases when shorting this lead to one of the others, the unpaired third lead is the transistor base. The lead to which the base is shorted to cause an increase in resistance is the emitter.

With the ohmeter connected as above, shorting the third lead to one of the other two should produce an increase in resistance if the third lead is the base. If not, it will be necessary to repeat the procedure with the other pairs until the right combination is found. This method isn't foolproof and may not work for some transistors. In such cases the base still can be identified, but you may not be able to identify the emitter and collector.

If, when using the preceding technique, the base can't be identified; and if you notice a high resistance in both directions for a given pair of leads, then most likely the third lead is the base. This can be checked by assuming this lead is the base and connecting the ohmeter between it and the other two leads to determine if the transistor is pnp or npn. If you can successfully determine transistor polarity, then you have correctly identified the base lead. The collector and emitter can be identified by reversing the two unknown leads during one of the grading tests.

The rough grading of transistors is accomplished using three simple circuits that will separate the transistors into four general categories:

1. Audio transistors with a beta (dc current gain) of less than 120.

2. Audio transistors with a beta of 120 or greater.

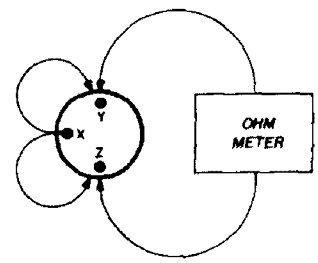



**fig. 3. Identifying transistor leads. Leads are labeled as shown; ohmeter leads are alternated to obtain lowest reading. The third lead, x, is shorted to one of the other transistor leads. If resistance increases, x is the base lead. The lead to which the base is shorted to obtain a resistance increase is the emitter; remaining lead is the collector.**

3. Transistors usable at radio frequencies.

4. Transistors not gradeable by these methods.

Be sure in all tests to observe proper battery polarity for pnp and npn, or erroneous results will be obtained.

## checking transistor beta

The first step in grading transistors is to determine their approximate beta. The circuits of figs. 4 and 5 are used. Connect the circuits together; then place two known transistors, such as the HEP251, into the circuits. Note that transistor sockets must be used.

Connect the battery and either a pair of headphones or a scope, as shown in fig. 5. Adjust R1 until oscillation occurs, indicated by a tone in the headphones or a trace on the scope. Now substitute the transistors to be tested, one at a time,

into the circuit of **fig. 4** and note whether oscillation occurs. If such is the case, the transistor has a beta of 120 or more and can be placed in the high-beta category. The twin-tee oscillator in this circuit will not oscillate with transistors having a beta less than 120. If the circuit won't oscillate, save the transistors for use in the remaining two tests.

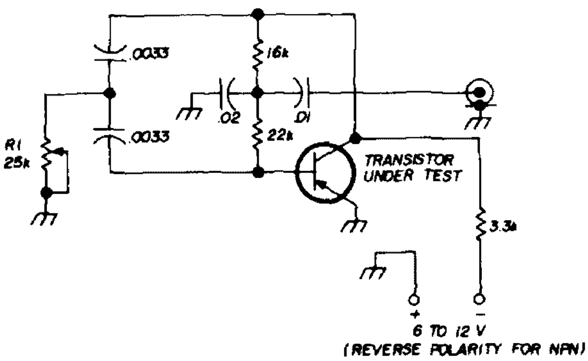



**fig. 4. A twin-tee oscillator used to determine beta of unknown transistors.**

## identifying emitter and collector

Suppose you were able to identify only the base lead during the polarity tests described earlier. The unknown leads should now be reversed, while keeping the base connection unchanged, until the circuit oscillates. If oscillation still does not occur, then again save the transistor for the remaining two tests and reverse the leads in these tests until results are obtained.

Don't hesitate to reverse the leads, as it's unlikely that you'll zap the transistor. Even if you should burn out an occasional transistor, you've lost nothing because an unidentifiable transistor is of no value anyway. Note that in some cases during this test it may be necessary to vary R1 until the circuit oscillates.

The transistors in this category are to be considered as having a high beta and may be used in circuits requiring such a device.

## amplification test

Transistors that fail during the high-beta test should be tested to see if they will amplify. To perform this test, set up the circuits in **figs. 4** and **5**. If possible, use a scope instead of headphones, as additional information can be obtained.

Replace the known transistor in the circuit of **fig. 5** with the transistor to be tested, observing the correct battery polarity. Note whether amplification occurs, as indicated by an increase in the tone in the headphones or a change on the scope trace. If amplification occurs, the transistor is good and is in the low-beta category.

## oscilloscope clues

When using a scope for these tests, note whether part of the waveform is distorted or clipped. If this is the case, then most likely the transistor is in the large-signal class or in a class that can't be properly graded by these techniques.

If the scope reveals no amplification whatever, even badly distorted amplification, the transistor is probably bad or is in the category "ungradeable by these methods." Transistors in this low-beta

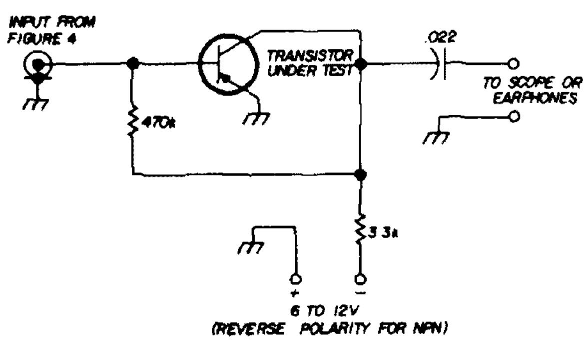



**fig. 5. Circuit used with that of fig. 4 in the amplification test.**

class can be used for most general applications in which a low-beta device is acceptable.

## rf test

Any of the transistors in the previous cases can be given the rf test, regardless of whether they passed or failed any of the earlier tests.

For the rf test, use the circuit shown

in fig. 6. Choose a crystal within the range of your receiver. Choose the constants of C1, L1 to resonate within the appropriate range. Use a 2N706 or equivalent for Q. Note that battery polarity is important for the device chosen.

Adjust C1 until oscillation occurs. Set C1 midrange between the two points where oscillation ceases. If a receiver isn't available, a grid-dip oscillator set to the diode position can be coupled to L1. I've shown circuit constants in fig. 6 for the ranges in which I separate my transistors: up to 30 MHz and up to 50 MHz. By using the proper crystal and circuit constants, transistors can be graded into any desired radio-frequency range.

Once the circuit is oscillating with the 2N706, simply plug in the transistors to be tested, one at a time, and note whether oscillation occurs. If the circuit oscillates, you know the transistor is good up to the frequency of the crystal. It shouldn't be necessary to reset C1 for different transistors. If no oscillation occurs, the transistor is not an rf type.

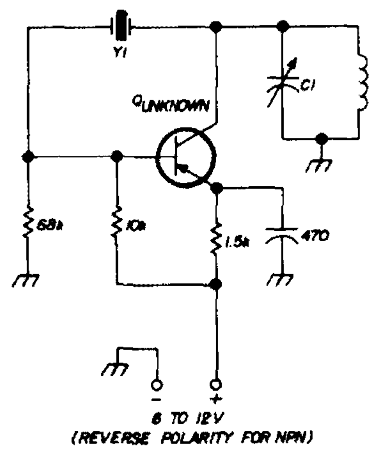
## power, voltage, and current

It's very difficult without manufacturers' specifications to determine power, voltage, and current ratings of unknown transistors. The assumption I make is that all the transistors *will* operate at 12 volts minimum and will dissipate at least 60 mW. So far, this has been a safe assumption, because by keeping within these ratings I haven't burned out any devices. As for the current rating, I usually assume that if the device is kept within its power rating, the current rating will take care of itself.

If you need transistors with a higher power rating, and you have several devices of the same type, then by all means try out a sample in your circuit to determine if it gets too warm. I don't know of any other approach to this problem.

## using graded transistors

At this point you may question the logic of separating the transistors into only three groups. The general feeling among most experimenters is that transistors with different numbers can't be readily interchanged. This is not usually the case. I've found from experimentation that, if a circuit calls for a transistor with high beta, in about 90 percent of the cases one of the transistors I've graded as




**L1 6 turns no. 18, ½" diameter, 3/8" long**

**Y1 25 to 30 MHz overtone crystal (C1 = two 7-45 pF trimmers in parallel)**

**50 to 60 MHz overtone crystal (C1 = 3-12 pF trimmer)**

**fig. 6. Circuit for determining if unknown transistor will work at radio frequencies.**

high beta will work well. In the majority of cases, the first transistor I choose from the group will work well.

## applications

Once your transistors have been graded into the general categories, they're ready for use in new projects. When building a new circuit, be sure that transistor sockets are used so substitution can be made easily.

To use your graded transistors, first look at the original transistor in the circuit. From its specifications, determine if the transistor is high beta, low beta, or general rf. Keep in mind we're talking about small-signal devices — those of less than 150 mW dissipation.

Once the category of the transistor has

been determined, select samples from your graded stock and substitute them into the circuit, one at a time, until optimum performance is obtained. In most cases, the first or second transistor will work.

As an example, suppose your circuit calls for a 2N217, which has a high beta, pnp polarity, and a dissipation of about 150 mW. This fits into our category of high-beta devices, but we previously made the assumption that all our units would dissipate about 60 mW. In this case, choose a transistor that's duplicated several times in the high-beta category. Substitute one of the high-beta graded units into the circuit and see if it works. If such is the case, leave the transistor in the circuit for five or ten minutes. (Make certain your power supply is fused.) If the device burns out or gets hot quickly, then try another variety.

While this method of determining transistor dissipation may not seem very scientific, I'm at a loss for a better method. I've tried to guess transistor dissipation from the case size, but I've found that some devices with large cases will dissipate little power, while others with small cases will dissipate quite a bit.

## rf transistors

The rf test described above is not a complete and thorough test. It will only test whether a transistor will oscillate at a specific frequency or below. In practice, transistors in this class will work well as oscillators, multipliers, and amplifiers in noncritical applications such as transmitters. In receivers, this class of transistor may be too noisy. This criteria may be easily determined by substitution.

## conclusion

The techniques described in this article are by no means intended for precise determination of transistor parameters. Rather, the techniques are presented so you can classify unidentified devices into several general categories. These categories can then be used as a starting point for building many transistor projects.

ham radio

# harmonics, distortion and splatter

A brief review of the reasons behind nonlinear amplifier operation

A. E. McGee, Jr., K5LLI, 2815 Materhorn Drive, Dallas, Texas 75228

Judging by what I hear, and even occasionally by what I read, there seems to be a lack of understanding concerning harmonics and other associated phenomena. Every ham knows the definition of a harmonic; that is, a harmonic is an exact whole-number multiple (2 times, 3 times, etc.) of a given fundamental frequency. Some think, however, that harmonics are a part of the natural order of things and that all signals will have harmonics. Others consider harmonics to be quite *unnatural* and that their presence is a sure sign of improper design or operation of a circuit. Neither of these assumptions is entirely true, as you will see later on.

## harmonic-free signals

There is one, and only one, type of signal completely free of harmonics, and that is the perfect sinusoid or sine wave. A sine wave is the wave shape formed by rotating a loop of wire at a constant speed in a uniform magnetic field. It's called a sine wave, because the instantaneous output voltage is proportional to the sine (a value found in a set of trigonometry tables) of the angle of rotation. The sine wave and how it is derived is shown in **fig. 1**.

A sine wave can be generated electronically at audio and radio frequencies by specially designed circuits. Just any oscillator circuit won't necessarily produce a sine-wave output, although many will approximate it closely. No matter how you produce it, a signal with a true sine-wave shape will have only one frequency, and harmonics will be nonexistent.

## distortion produces harmonics

Any variation from a true sine-wave shape causes other frequencies to appear along with the fundamental or lowest frequency. These extra frequencies will be harmonics of the original frequency, and their number and strength is determined by the shape of the wave. You could also say that the shape of the wave is determined by the number, strength, and phase of the harmonics present. The result is the same in either case.

The harmonics of a distorted sine wave are all perfect sine waves. The fundamental frequency is also a perfect sine wave. This fact is rather confusing and hard to grasp at first, but it can be more easily understood if you will realize that the wave shape you see on an oscilloscope is a composite of all the frequencies present in the wave. The instantaneous voltages of the harmonics add to and subtract from the instantaneous voltage of the fundamental sine wave to give the resultant wave shape. See fig. 2.

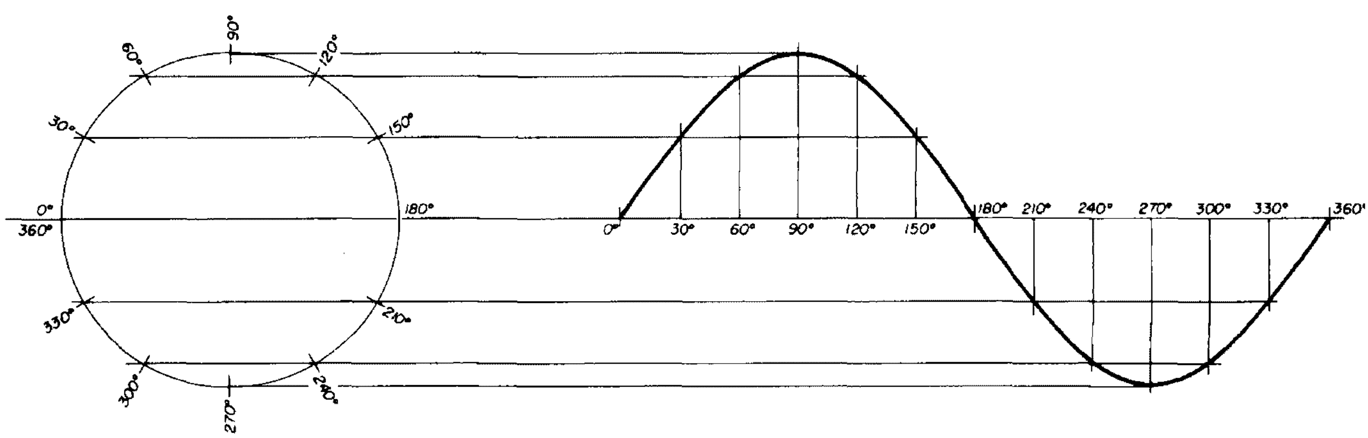



**fig. 1. The sine wave; the only wave that has no harmonic content.**

The total number of harmonics present in a wave can range from one, as in a sine wave with some second- or third-harmonic distortion, to thousands, as in a good square wave.

Many wave shapes also have a dc component. This is caused by the part of the wave on one side of zero being larger than that on the other side. As a general rule, it can be said that only even harmonics will produce a dc component.

## mathematical proof

I have made a number of statements without giving proof that any are true. It can be proved mathematically, by means of a Fourier analysis, that a distorted or nonsinusoidal electrical wave is the sum of a fundamental sine wave, a series of sinusoidal harmonics, and a dc or average value. I'll leave this to the mathematicians, and take their word for its truth.

## causes of distortion

If you generate a pure sine-wave signal, then feed it through an amplifier to increase its strength, chances are you will no longer have a pure sine wave at the output of the amplifier. This is because no vacuum tube or transistor is a perfectly linear device. That is, the output current doesn't increase and decrease exactly in proportion to an increase or decrease in input voltage or current. Many amplifiers, however, will be very nearly linear over a certain range of input voltges; and by using negative feedback or other special techniques, an extremely good amplifier can be built. I just wish to emphasize that unless care is taken, there will be harmonic frequencies present in the output of an amplifier that were not present at the input.

Distortion in an amplifier isn't always a bad thing, and may be purposely introduced, as in the case of a frequency-multiplier circuit. A frequency multiplier is usually biased so that output current flows during only a small portion of the input cycle. This distorts the output signal greatly and causes the generation of strong harmonics.

All class-C amplifiers produce a great deal of distortion and many harmonics. Class-C amplifiers are used only to amplify cw or fm signals, however, and the tuned-output circuits attenuate the harmonics to a reasonable level.

## harmonics and splatter

When more than one signal is present in a nonlinear circuit, the situation becomes more complicated. Not only will harmonics of the signals be present, but the nonlinear amplifier will also act as a mixer, and the sum and difference frequencies of the fundamental and the various harmonics will be present in the output. For example, let's see what would happen if two sine-wave signals were fed into an amplifier where considerable distortion is present. For convenient numbers I'll use 1000 and 1001 kHz as the two frequencies. This would be the equivalent of a two-tone single-sideband suppressed-carrier signal.

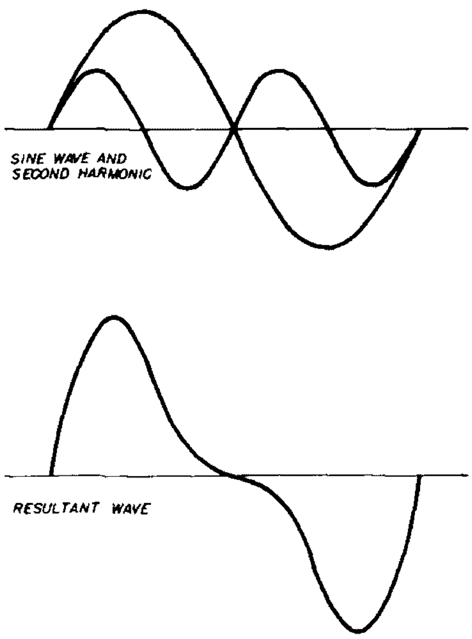



**fig. 2. A distorted wave consisting of a fundamental sine wave and a large second harmonic.**

The harmonics and the sum frequencies generated by the amplifier would be greatly reduced in strength by the selectivity of the tuned-output circuit. However, some of the difference frequencies would fall within the passband of the tuned circuit. The second harmonic of 1001 kHz (2002 kHz) will mix with the 1000-kHz signal to get an output of 1002 kHz; the second harmonic of 1000 kHz (2000 kHz) will mix with the 1001-kHz signal to get an output of 999 kHz; the third harmonic of 1001 kHz (3003 kHz) will mix with the second harmonic of 1000 kHz (2000 kHz) to get an output of 1003 kHz; the third harmonic of 1000 kHz (3000 kHz) will mix with the second harmonic of 1001 kHz (2002 kHz) to get an output of 998 kHz; and so forth, depending on the amount of distortion in the amplifier and the number of harmonics present.

You can see that although we started out with only two signals, at 1000 and 1001 kHz, we now have six signals; at 998, 999, 1000, 1001, 1002, and 1003 kHz. These extra signals are caused by distorting the wave shapes of the original signals by amplifying them with a nonlinear amplifier. With a properly designed and operated amplifier, these extra signals will be so much weaker than the desired signals that they will be of no consequence.

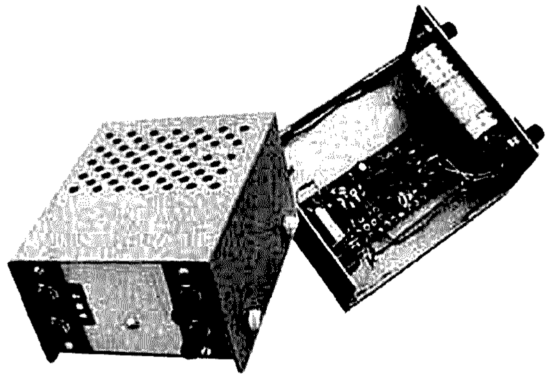
However, even the best amplifier, when over-driven, badly loaded, or otherwise improperly operated, will put out a great many strong undesired signals. This is why it's important that linear amplifiers be correctly operated, or else they may suddenly become nonlinear amplifiers, and can easily put out spurious signals (known as splatter) covering an entire amateur band.

In conclusion, the sine wave is the only wave form that has no harmonic content, any distortion of a sine wave means that harmonic frequencies are present, and when two or more signals are amplified by a nonlinear amplifier, a great many unwanted signals may appear in the output.

**bibliography**

1. A. E. Richmond, "Calculus for Electronics," McGraw-Hill Book Co., Inc., 1958.
2. Nelson M. Cooke, "Mathematics for Electricians and Radiomen," McGraw-Hill Book Co., Inc., 1942.
3. Irving Gottlieb, "Basic Pulses," John F. Rider Publisher, Inc., 1958.
4. "Single Sideband for the Radio Amateur," The American Radio Relay League, Inc., 1954.

**ham radio**



# improved superregenerative receiver

Atuyuki Iwakami, JA1BHG, 2-203, 1-3 Kasumigaoka, Fukuoka, Iruma, Saitama, Japan

The superregenerative receiver is a simple low-cost system often used in portable equipment operating on the amateur vhf and uhf bands. Although this circuit offers high sensitivity, it suffers from poor selectivity, high noise level, oscillator radiation and hangover. Hangover results in blocking that limits sensitivity because the receiver is swamped by its own residual signal.

The high sensitivity of the superregenerative detector is due to the use of an alternating quench voltage, usually between 20 and 300 kHz. The regeneration control is set so the detector goes into oscillation on each positive peak of the quench voltage: on each negative swing the oscillator is cut off.

The superregeneration principle can be applied to any oscillator circuit; a grounded base Hartley circuit is shown in **fig. 1.** If the bias is gradually increased with the "regeneration" control the circuit will break into oscillation. When the amplitude of oscillations overcomes the base bias voltage on the negative portion of the swing, rectified current through the base-emitter diode charges capacitor $C_b$, putting a negative bias on the base that runs off the transistor. When $C_b$ has discharged through resistor $R_b$ the circuit begins to oscillate again. Hence, the quenching frequency is determined by the base-bias voltage and the time constant of $C_bR_b$. In **fig. 1** $R_b$ is the effective resistance from base to ground and consists of R1 in parallel with R2 and part of R3.

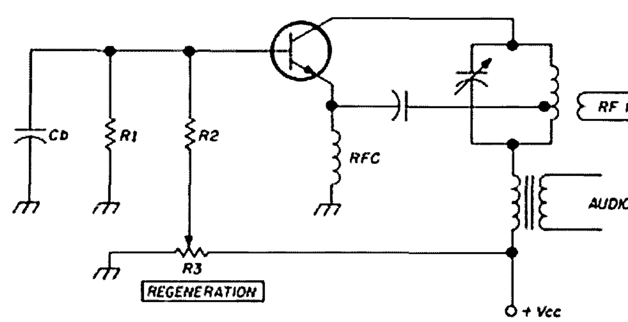



**fig. 1. Simple superregenerative circuit using the grounded-base Hartley circuit.**

An analysis of one cycle of quenching is shown in fig. 2. The oscillation is triggered by the incoming signal and builds until it overcomes the bias potential, then is quenched. When there is no input signal, internal circuit noise acts as

This article originally appeared in the Japanese amateur publication, *CQ-Ham Radio,* and is presented here through the cooperation of JA1BHG, editor.

the trigger with a resultant "hiss" in the audio output. Hangover is due to the fly-wheel effect of the tank circuit, and the higher the Q, the more troublesome the hangover.

In a recent article K2ZSQ suggested a simple circuit addition to limit the affects of radiation and hangover.[1] The circuit change consists of adding a germanium diode across the tank circuit as shown in fig. 3. With this diode undesired energy is immediately dissipated after the oscillation burst. This eliminates hangover effects during the remaining period of the burst. Radiation is also lessened because the damping action of the diode lowers amplitude and shortens the duration of radiated pulse.

## practical receiver

The modified superregenerative receiver shown in fig. 3 has been used on 50 and 144 MHz with similar results. The added diode is not critical, and any germanium point-contact seems to work well. The transistor should be a vhf type for best operation, but this isn't too critical either. I used Japanese types 2SC372 ($f_T$ = 150 MHz) and 2SC387 ($f_T$ = 900 MHz), but Motorola equivalent types HEP55 and HEP56 will perform as well.

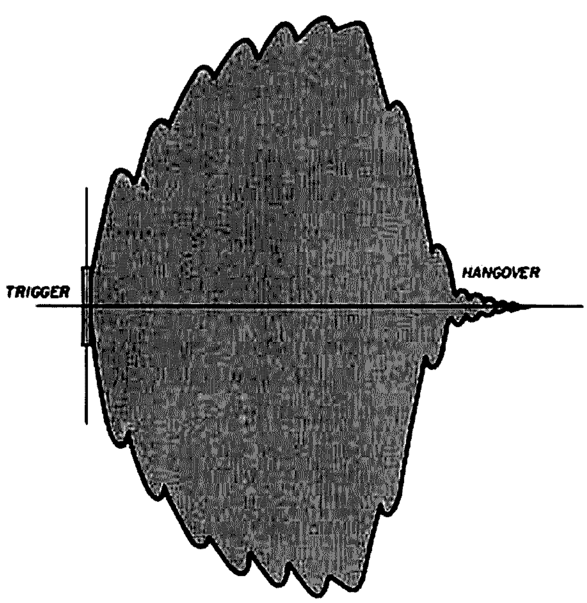
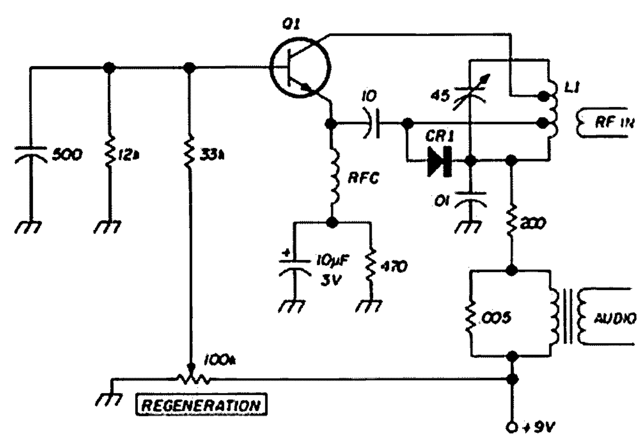



fig. 2. One cycle of superregenerative operation shows the input trigger, oscillation, quenching and hangover.

Smoother operation is obtained by tapping the collector down on the final tank coil as shown in fig. 3. Oscillation can also be improved by moving the emitter tap down, but effective damping action requires the emitter to be tapped as high as possible on the coil.




**50 MHz**

L1 7 turns no. 20, 9/16" diameter, 3/4" long. Collector tap at 2½ turns, emitter tap ½ turn from ground

Q1 HEP 55

RFC 10 $\mu$H

**144 MHz**

L1 4 turns no. 16, 3/8" diameter, 9/16" long. Collector tap at 1½ turns, emitter tap at ½ turn from ground

Q1 HEP 56

RFC 20 turns no. 26, closewound on ¼" form

fig. 3. Practical vhf receiver for 6 and 2 meters using the modified superregenerative circuit.

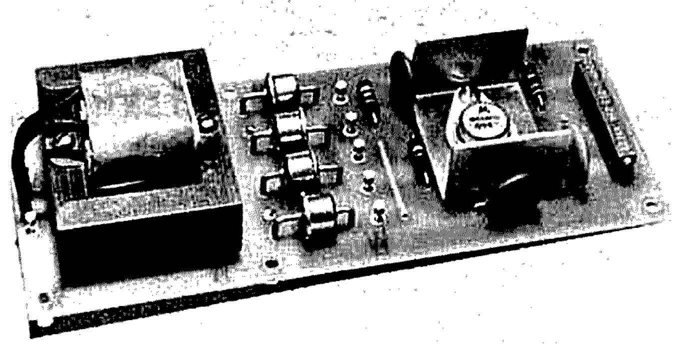
## results

Since this circuit is basically an oscillator, radiation was decreased very little by the addition of the germanium diode. However, the hangover effect was considerably improved. No noticeable decrease in sensitivity was found.


reference

1. Robert M. Brown, K2ZSQ, "No-Radiation, No-hangover 28-MHz Superregen Receiver," *ham radio*, November, 1968, p. 70.


ham radio



# a flexible voltage-regulated power supply

Putting the Motorola MC1460R regulator IC to work in a useful circuit for the experimenter

Ron Vaceluke, W9SEK 17 W. 540 Hillcrest, Wood Dale, Illinois 60191

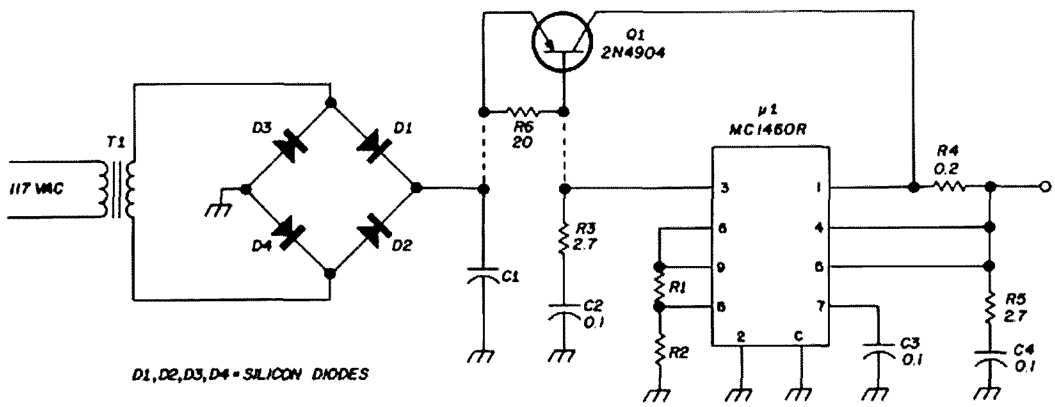
The power supply described here was originally designed for a 2-meter fm receiver.[1] However, it can be used for any equipment requiring a regulated low-voltage power supply.

The integrated circuit is a Motorola MC1460R, which can handle up to 20 volts input at 600 mA maximum. For higher voltages, the MC1461R (35 V), or the MC1561R (40 V) can be used. The case is a small diamond shape, which can be mounted in a heat sink.

## the circuit

The power supply requires a minimum of parts (fig. 1). Resistors R1, R2 constitute a voltage divider that determines output voltage. With R2 = 6.8k, R1 will be $(2V_O\text{-}7)$k ohms, where $V_O$ is the output voltage. R1 can be fixed or variable. Resistor R4 is the current-limiting resistor, which determines the short-circuit load current. If an external pass transistor is not used, values for R4 may be obtained from fig. 2 for various short-circuit load currents.

Since the IC transistors have high frequency capabilities, there's a chance of oscillation with this device; therefore, some means must be used to suppress this tendency. Networks composed of R3, C2 and R5, C4 form suppressors for input and output respectively.




C1 filter capacitor (value depends on current drawn from supply)

R1, R2 see text

R4 0.2 ohms (for use with Q1 at 2A maximum output)

T1 12 VAC, .45A (Stancor P-8392)

**fig. 1.** Schematic of the regulated supply. If an external pass transistor isn't used, terminals E and C are jumpered. Components outside the dashed line are mounted externally.

## construction

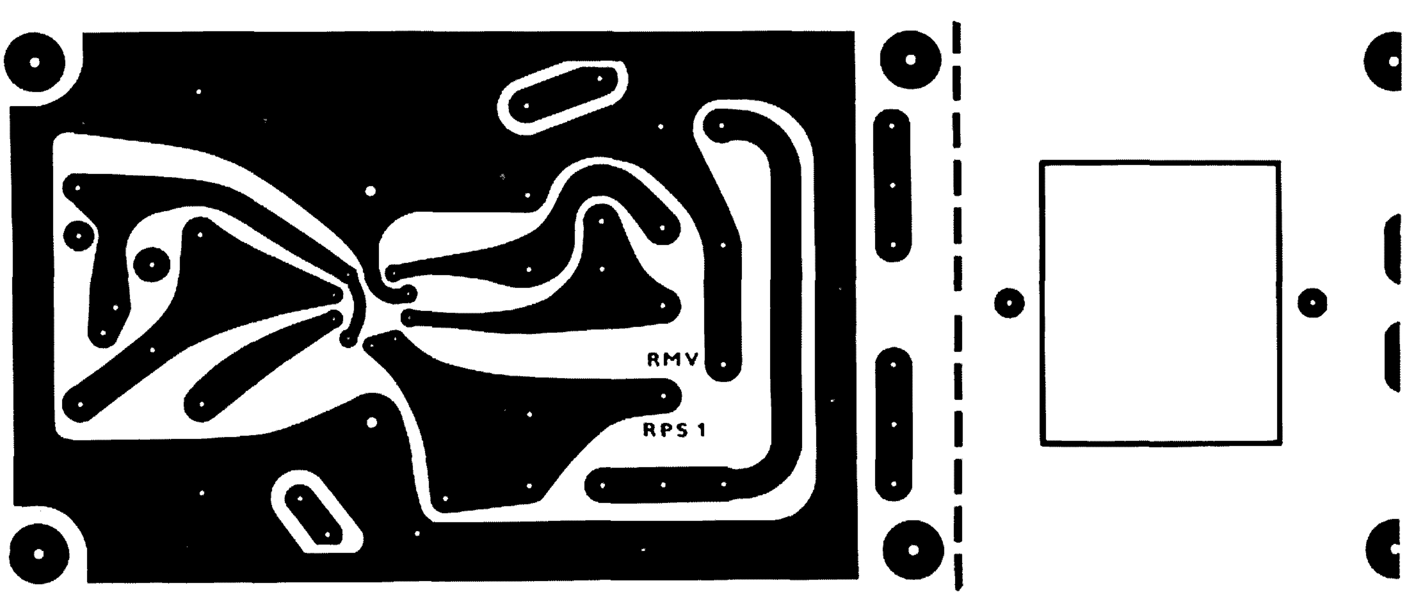
Because of the vhf transistors in the IC, care must be used in wiring the circuit. Hand wiring can be used, but it's not recommended unless vhf-type construction techniques are used. This means extremely short leads and vhf grounding methods. A printed-circuit board is the best solution. A PC board can be made from the template shown, or it can be purchased.*

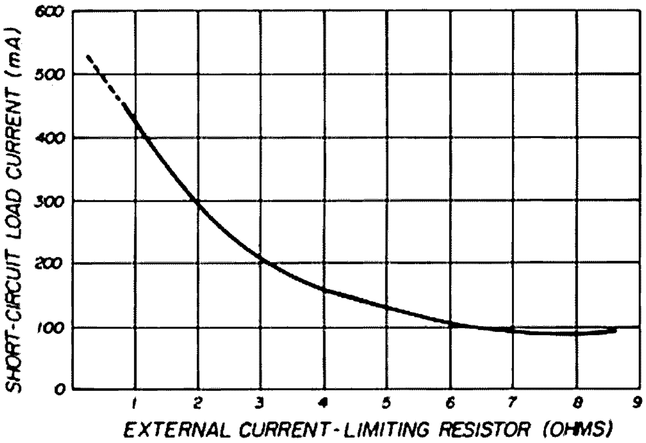
The PC-board layout has a certain amount of built-in flexibility. A small power transformer can be included on the board or the board can be cut on the dashed line if a larger transformer is desired.

Space is provided for four separate rectifier diodes rather than a diode assembly, so junk-box components can be used. The circuit board is arranged so that three

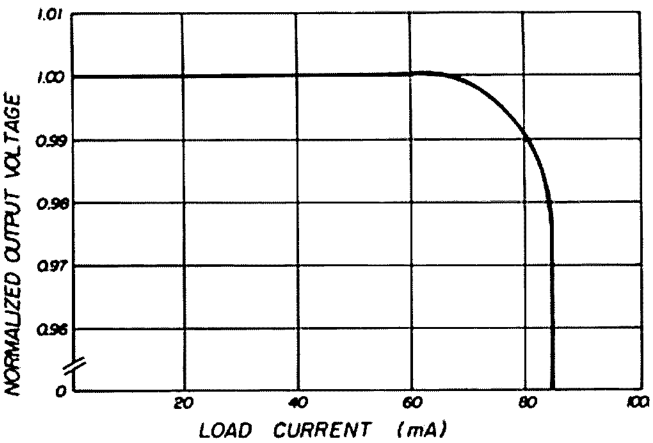
*A G-10 epoxy board, drilled and tinned with staked terminals, is available from RMV Electronics, P. O. Box 283, Wood Dale, Illinois 60191. $3.25 each postpaid.

**fig. 3.** Full-size PC board template. Board can be cut on dashed line if the transformer is mounted externally.






A




B

**fig. 2. Curve A is used to determine the value of R4 if a pass transistor isn't used; curve B shows voltage and current when R4 is 6.8 ohms.**

different styles of potentiometer can be used for R1. A fixed rather than a variable resistor may be used if desired.

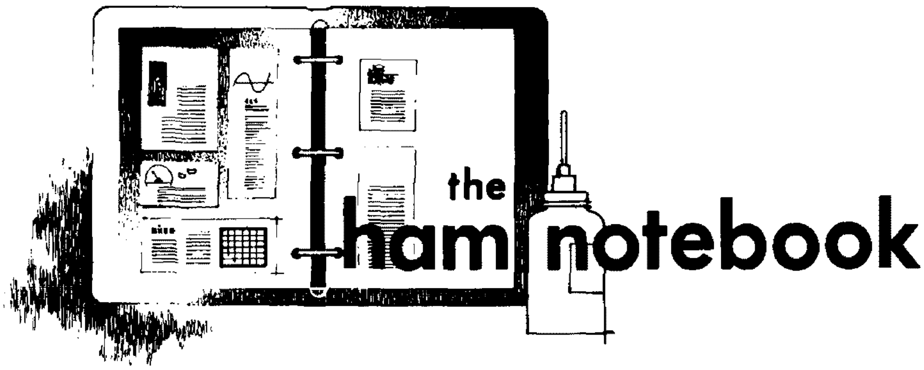
The heat sink can be made from a piece of aluminum with a ½-inch hole to pass the IC leads. Make sure the leads don't touch the heat sink. The size of the heat sink will depend on the power to be dissipated.

Examination of the photo shows a jumper, which should replace R6 if an external pass transistor is not used. The only other external component is the filter capacitor, C1.

reference


1. R. M. Vaceluke, W9SEK, and J. C. Price, WA9CGZ, "Vhf Fm Receiver," *ham radio*, September, 1970, p. 22.


ham radio




## frequency-sensitive resistors

Deposited-film resistors operating in circuits at hf aren't always to be trusted if precise resistance values are needed.

While at work, I needed a fairly husky precision resistor for a phase-sensitive network in a 100-kW transmitter. When I learned that the resistor order would be delivered in the distant future, I decided to check our stock of deposited-film resistors to see if I could use an acceptable substitute. Much to my surprise, I found that at frequencies as low as 14 MHz most values were completely inadequate.

The resistors were checked on an rf bridge. I found that the "r" component generally tended to run quite low – at least 10% low in a 1% resistor – except that values in the 25k and higher ranges showed violently reversed characteristics. Resistors from some manufacturers were fairly accurate in this frequency range, but many showed bulk resistance effects. For example, a 25k unit would shoot up to 100k or more!

I haven't completely checked out the results of these tests, but it appears that values of about 300 ohms or so tended to be capacitive at these frequencies, which is opposite to what I would have suspected. Units of 200 ohms or less tended to be slightly inductive. So you might bear this in mind the next time you need precision resistors in a frequency-sensitive application.

**Bill Wildenhein, W8YFB**

## fm repeater receiver isolation

If you are interested in obtaining channel isolation in your fm-repeater receiver, the Motrac units made by Motorola are worth considering. These units have a five-cavity front end and are readily available from surplus sources dealing in fm gear.

## continuous tuning for fm converters

The Collins 75A-3 and 75A-4 receivers, with their calibrated-tuning capability, can be used with most crystal-controlled converters to provide continuous coverage of the two-meter fm band.

The appropriate converter is one having an output between 26-30 MHz. The 75A-3 receiver will tune this range directly; the 75A-4 will cover most of this range, and an appropriate crystal can be selected to cover the frequency portion desired.

On all later production models of the 75A-3 (and on all 75A-4 receivers) a

broadband i-f output jack is provided, which was originally intended for use with a panadapter. This output can be coupled via 52-ohm coax to the i-f strip of any of the surplus Motorola receivers, most of which use the same 455-kHz i-f.

For early 75A-3 receivers that don't have the panadapter output, the instruction manual gives complete details for installing the output jack, which is a very simple job.

If you have a 75A receiver, this is an excellent way to become acquainted with the two-meter band and discover which spot frequencies are active. Then you can decide which crystals to purchase for the fixed-tuned equipment you would eventually use. You can't beat fm for communications reliability and noise-free operation.

# inexpensive WWV converter

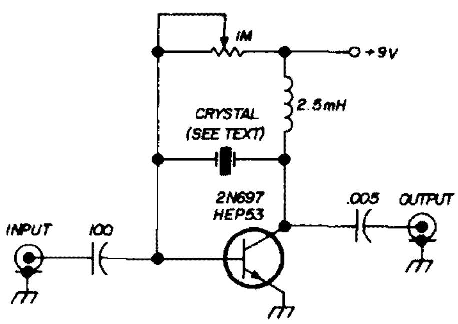



**fig. 1. Simple mixer circuit for converting WWV signals to amateur bands.**

Here's a simple circuit that will convert WWV signals above 5 MHz to amateur bands between 80 and 15 meters. The converter consists of a simple crystal mixer using a single npn transistor (fig. 1).

A variety of crystal frequencies can be used, depending on which WWV signal is desired. See table 1. The exact output frequency may be determined by finding the difference or sum of the desired WWV frequency and the crystal frequency.

No tuned circuits are used, so other

**table 1. Crystal frequencies that can be used to convert WWV to the various amateur bands.**

| ham band | WWV frequency (MHz) 10 | 15 | 20 | 25 |
|---|---|---|---|---|
| 80 | 6-6.5 | — | — | — |
| 40 | 2.7-3.0 | 8-7.70 | — | — |
| 20 | 4-4.35 | — | 6-6.65 | — |
| 15 | — | 6-6.45 | — | 3.55-4.00 |

signals will be received in additon to WWV. For example, if a 6.5-MHz crystal were used to convert WWV at 10 MHz to 3.5 MHz, all signals on 3.5, 3.0, and 10 MHz would be picked up. However, WWV is usually strong enough to be picked out easily. If necessary, tuned circuits could easily be installed between the antenna and the converter.

WWV transmits much interesting data in addition to time ticks. Schedules and types of information are published in the ARRL handbook.

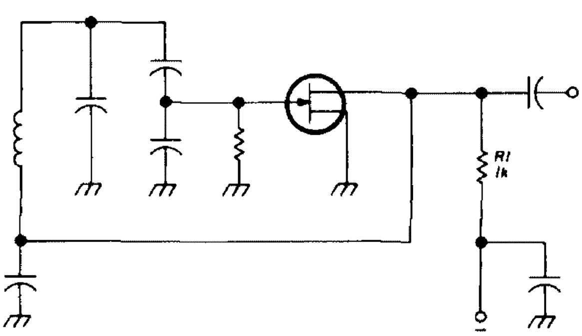
Doug Pongrance, WA3JBN

# cure for cranky oscillators

In all the Vackar oscillators I've built using a Motorola MPF102 jfet, I've found that the usual circuit (fig. 2) is reluctant to oscillate with low drain voltage. A quick and sure fix is to replace the usual 1k-ohm drain load resistor, R1, with a small rf choke. I found that video peaking coils pulled from an old TV set are adequate. With such a modification, the circuit will oscillate vigorously in the 8-12 V power-supply range.

Bill Wildenhein, W8YFB

**fig. 2. Typical Vackar oscillator using the MPF102 fet. Replacing R1 with a small rf choke ensures oscillation with low drain voltage.**



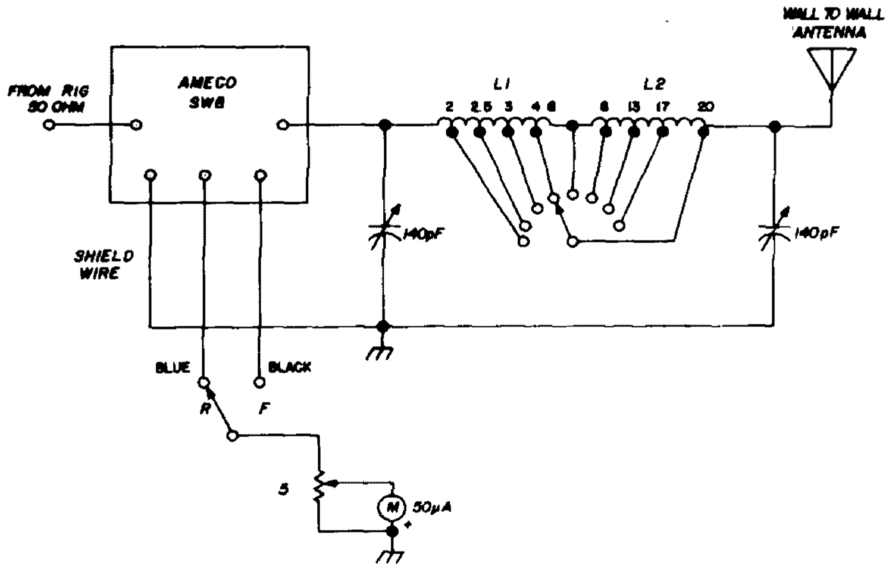



fig. 3. Bridge and tuner for random length antennas. Variable capacitors are Hammarlund type MC140. L1 is 8 turns per inch; L2 is 16 turns per inch. Both coils are 1½-inch diameter. L1 is tapped at 2, 2½, 3 and 4 turns; L2 at 8, 13 and 17 turns.

# wall-to-wall antenna tuner

Although a well-elevated outdoor antenna is best for radio communications, such an antenna is often impractical or impossible to erect (due, for example, to space restrictions or a grouchy landlord). Fortunately, communications on the ham bands can be effective with only a simple short-wire antenna installed indoors.

I have contacted many stations on the West Coast and in the Midwest using only a horizontal wire strung between two walls. However, the antenna must be properly terminated to be effective. The terminating impedance for most rigs is 50 ohms.

All the equipment needed to terminate the antenna properly is shown in fig. 3. In addition to the rf input and output terminals, the Ameco Model SWB bridge has three leads, which should be connected as shown. The switch selects either forward or reverse power. We want as high a reading as possible in the forward position and as low a reading as possible in the reverse direction. The antenna tuner is adjusted for optimum impedance match between transmitter and antenna.

## construction

For convenience, I mounted my bridge in a 4 x 4 x 2-inch aluminum box. The Ameco bridge, which comes with Amphenol SO-329 connectors, is 4½ inches long. It won't fit into a 4-inch-long box, so here's what to do.

Remove the front and rear panels of the box. Then saw one edge of the box so you can spread it apart to accommodate the bridge. Drill holes to allow the Amphenol connectors to protrude. Use a ¾-inch circle cutter.

Mount the bridge on the bottom of the box with screws. Close the box, using a small angle bracket and pk screws to keep it closed tightly.

The 4-inch box also includes the meter (1½-inch square), a miniature toggle switch, and a miniature potentiometer. There's plenty of space for these components if they're mounted on the front panel.

## the tuner

The tuner is housed in a metal box, 8 x3 x 2-3/4 inches. The variable capacitors are Hammarlund type MC140. The inductance consists of two separate coils in series, each 1½ inches in diameter. The coils are mounted side-by-side, about one half inch apart.

## tune-up

If you have a 50-ohm dummy load, adjust your rig for optimum output. Position the switch to R, and select a tap on the tuner that gives minimum reading on the meter. Adjust the variable capacitors for a minimum meter indication.

I use this system with an indoor antenna on 15 and 20 meters, my favorite bands. The tuner and bridge will probably match any random-length antenna. I've used the circuit of fig. 3 with a vertical whip, 6 feet high, and with an outdoor long-wire antenna 50 feet long. Results have been very good.

I. Queen, W2OUX

# plastic protective material

A self-adhesive plastic sheet product is on the market that is of interest to amateurs. The material I use is called "PLAIN-VU" manufactured by Carr Adhesive Products, Inc., Somerville, Massachusetts.

PLAIN-VU is a clear plastic contact material, which is good for covering homemade dials, panels, I. D. cards, licenses, QSL cards, and the like. It's available in stationery stores in sheets 9 x 11-5/8 inches, two sheets for about a dollar.

Paul White, W6BKX

# fm deviation meter

With fm becoming more and more popular, a deviation meter is a nice piece of equipment to have in your station. The method described can be used for most frequencies, and on 2 meters it works fine.

## features

The deviation meter allows you to check deviation of on-the-air signals as well as your own. The use of an oscilloscope is ideal, and peak deviation can be monitored.

A two-channel fm receiver is used, with channel A tuned to the desired frequency and channel B tuned either 15 kHz above or below channel A. The setup is shown in fig. 4. Discriminator output is fed to the vertical input of a scope. Internal sweep is used for the horizontal plates.

With no signal, noise will be displayed on the scope. When an unmodulated signal is received, a straight line will appear, possibly with some noise, depending on signal strength.

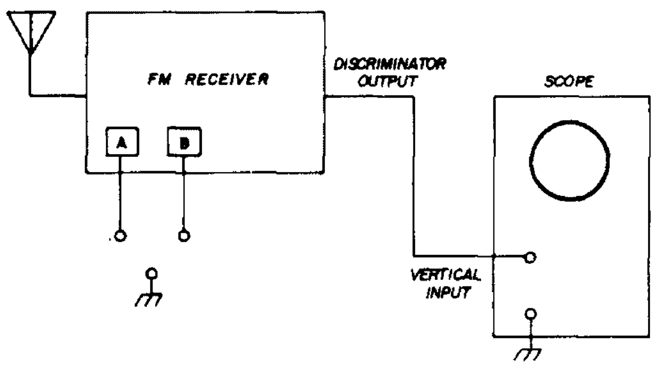



**fig. 4. Fm deviation meter using 2-channel receiver and oscilloscope. Method allows monitoring of peak deviation of received signal or your own.**

## operation

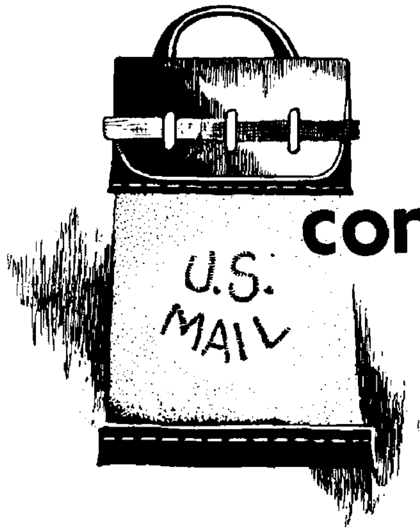
Switch the receiver to the desired signal.

Adjust vertical centering so that the line appears in the center of the display. Now switch to channel B. The line will swing either up or down. Calibrate the scope vertical gain for a reasonable display.* Switch back to channel A, and modulate the transmitter to be tested. If voice peaks hit the calibration points, the signal is deviating ±15 kHz.

A more elaborate unit could be made using a three-channel receiver, which would allow calibration above and below the received frequency. Two channels should suffice, however, since the signal deviates equally well both ways.

Vern Epp, VE7ABK

*Or use a grease pencil to mark deviation limits on the face of the scope tube. editor.




# comments

## antenna dimensions

Dear HR:

In the *ham notebook* section of the June, 1970 issue the antenna dimensions presented by WA9JMY seem to have a discrepancy in the "inverted-vee" column.

It has always been my experience that if I had a dipole of a certain length, resonant at a certain frequency, and then positioned it in the inverted-vee configuration, the resonant frequency *increased.* Thus, to maintain resonance at the original frequency, I had to *increase* the length such that the length of the inverted vee at the original frequency was *greater* than the length of the dipole at the same frequency.

The figures given by WA9JMY do not seem to be in line with my experimental findings. Throughout the chart, the inverted-vee lengths are *less* than those for a resonant dipole at the same frequency. Realizing that there is a multiplicity of factors involved in antenna work, I took a peek at the *ARRL Antenna Book,* 11th edition, page 204, and it seems to confirm my results.

"Sloping of the wires results in an increase in the resonant frequency and a decrease in feedpoint impedance and bandwidth as the angle between the two wires is decreased. Thus, for the same frequency, the length of the dipole (used as an inverted vee) must be increased somewhat."

Has a mistake been made in the calculation of the lengths in the inverted-vee column?

Donald R. Nesbitt, K4BGF
Gainesville, Florida

*For each antenna length computed by the computer, I fed formulas that were taken out of the 1968 edition of the ARRL handbook. I did this with the assumption that they were correct and that no research into other books would be necessary. After receiving your letter questioning the correctness of the inverted-vee column I checked back into the handbook with a little more alertness. This is what I found: Quoting the last paragraph on page 350 of the 1968 edition:*

> *"When its ends are near the ground, the length of the wire in an inverted V antenna is slightly shorter than when the dipole is strung in a straight line. . . "*

*These are the formulas I fed to the computer:*

*Inverted vee: Length = 464/Freq.*

*Dipole: Length = 468/Freq.*

*These formulas indicate that the dipole will always be longer.*

*I did not have a chance to verify your source of information but it is evident that one of these books is wrong. If you can dig up any more info, I would appreciate you letting me know.*

James Barcz, WA9JMY

## inverted-vee length

Dear HR:

I have checked out both the Antenna Handbook and the Handbook and sure enough, they seem to be contradictory except for the qualifier: "When its ends are near the ground. . ." It makes me wonder just how near the ground "near" is!

As for any additional information which I can add, most of my inverted-vees have been close to a half-wave above ground with apex angles in the 140° to 120° range thereby placing the ends perhaps not "near the ground!!"

In an article by K4GSX, "Radiation Resistance of Inverted V Antennas," *QST,* October, 1968, using a ten-meter test antenna with an apex angle of 105°, the length as compared to the standard dipole varied from longer than the resonant dipole (determined by setting k = 0.95 in the formula L = 492(k)/f(MHz)) to shorter depending upon the height above ground. Experimentally, he measured values of k for an antenna with the forementioned delta and an L/D ratio of 230 (hardly typical!!) which were less than 0.95 for heights less than 0.3 wavelengths (ie: the inverted vee was shorter than the corresponding dipole) and values of k which were larger than 0.95 (ie: the inverted vee was longer than the corresponding dipole) for heights greater than 0.3 wavelengths.

Perhaps in this case the center height of about 0.3 wavelengths or less put the ends "near the ground . . ."

Lewis McCoy, W1ICP, in the July, 1968 issue of *QST* p. 42 suggested that as a starting point, the length given by L = 515/F(MHz) would be appropriate using the tried and true method of pruning from there. Antennas being the cantankerous beasts that they are this seems a wise suggestion!!!!!

I propose to run some checks on an experimental 40-meter setup which would be a typical amateur antenna. Number 14 or 16 wire used with the small glass end insulators fed with a half-wave of coax (electrical of course) and adjustable both in height and apex angle. I'm interested in checking this type of antenna since I think that it represents the way most of the inverted vees are constructed and used. Hopefully it will satisfy my curiosity about how near "near" is!

Donald R. Nesbitt, K4BGF
Gainesville, Florida

## variable crystal oscillators

Dear HR:

I want to thank you for the fine article on vxo design. K6BIJ's circuit of **fig.** 4 which achieves 50 kHz frequency shift is certainly most impressive. I shall have to give it a try in the near future.

For your information, the circuit shown in **fig. 1** was originally published in the February, 1965 issue of *73 Magazine* and was specifically developed for use with the OSCAR 3 satellite.

These circuits would seem to have a wide area of applicability which has been pretty much overlooked. Perhaps your article will lead to additional applications.

Even the narrow shift of circuit 1 could be of value as a simple no-holes modification of existing transceivers for net and schedule type operation. For example, in my Heath SB-300 and 400 it could be readily adapted by making the vxo for 5 to 5.5 MHz and plugging its output into the system in place of the internal LMO (variable frequency oscillator) and still have the flexibility to zero in on the exact net frequency.

Your circuit which features increased frequency swing would have even more uses. I am thinking of one particular application but I wonder if it would be acceptable to the FCC. What I have in mind is its use by Novices. This could give

them the capability of vfo operation while still meeting the specific regulation requiring that their transmitters be crystal controlled. With just a few crystals they could cover the entire band available to them. Similarly, they could modify existing commercial units by replacing the internal vfos with vxos for novice operations.

Harley C. Gabrielson, K6DS
La Mesa, California

## antenna tuners

Dear HR:

The article (in your May, 1970 issue) about "compatible vs incompatible" tuners has many hams wondering about their antenna couplers. Initial investigations here have borne out the points of the article. One interesting discovery—I can match my antenna across the shorted-out section of the antenna-tuner coil and get a different set of tuning conditions—proof that what W2WLR said is true, that power *is* circulating around in all those extraneous portions of the tuner rather than getting out to the antenna!

Ade Weiss, K8EEG
Meckling, South Dakota

## simple wwv receiver

Dear HR:

Amateurs who have ham-band-only receivers and need a converter to receive WWV should consider International Crystal's OX oscillator and MXX-1 mixer. I put these two units together, and with a 6000 kHz crystal beat the 10.000 MHz WWV signal into my receiver at 4000 kHz in the 80-meter band.

Using a double-pole double-throw switch to put the WWV signal into the mixer for zeroing my crystal calibrator, and then out again, makes for one of the easiest and economical approaches to this problem.

James W. Harrison, Jr., WB4TBX
Norfolk, Virginia

# new products

## vfm-fm radiotelephone

Stoner Communications has announced a new low-cost vhf-fm radiotelephone designed for multi-purpose applications. The model VHF-30, six-channel, 25-watt unit features a mosfet front end, ic amplifiers and selective crystal filter to provide low-noise reception of weak signals. The equipment incorporates tropicalized epoxy board modules and splash and corrosion-proof construction. The unit can be used as a battery-operated base station; low current drain permits operation from any 12-volt source in base or vehicle applications. The VHF-30 covers the frequency range from 148 to 174 MHz with 0.6 $\mu$V sensitivity for 20-dB quieting.

For more information on the new VHF-30, use *check-off* on page 88, or write to Donald L. Stoner, Stoner Communications, Inc., 8751 Industrial Lane, Cucamonga, California 91730.

## cir-kit printed-circuit material

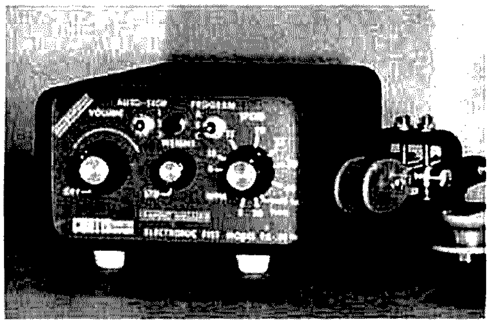
Cir-Kit is a revolutionary new material for the construction of experimental and prototype printed-circuit boards and consists of high-purity 0.002-inch copper foil tape – 1/8- or 1-16-inch wide – coated with a heat-resistant self adhesive and protected by lacquer. To use the material you simply remove the backing paper and place the adhesive side down in the desired position. Component leads are soldered to the Cir-Kit. If any circuit modifications are required it's a simple matter to remove the Cir-Kit material with a knife.

Cir-Kit can be bent, curved, used on both sides of a board to simplify layout or passed around the edges of a board to accomodate edge connectors.

The adhesive used on Cir-Kit actually increases its strength with age. The heat of soldering merely speeds up this process. The adhesive softens when it is heated but full adhesive strength returns as soon as the temperature drops to normal.

Rolls of Cir-Kit, 1/8 or 1/16 inch wide, 5-feet long are $.60 each; 100-foot spools are $9.95 each. Sheets of Cir-Kit, 6 x 12 inches are $2.50 each, 5 for $9.95. For more information, use *check-off* on page 88, or write to Cir-Kit, Box 592, Amherst, New Hampshire 03031.

# automatic cw identifier



Curtis Electro Devices has announced an advanced integrated-circuit keyer that incorporates an automatic identification-message generator in addition to the basic *Electronic-Fist* circuitry. In this new keyer, called the *Mnemonic,* a custom integrated-circuit read-only memory (ROM) provides permanent memory to generate the repetitive calls used by amateurs in normal and contest operation.

As an example, the operator might select any one of the three sequences below from a single ROM:

1. CQ CQ DE W1DTY K
2. CQ FD CQ FD DE W1DTY K
3. DE W1DTY K

Three auto-stop selections allow continuous cycling or a choice of two stopping points. The message can be transmitted once, continuously or every ten seconds.

In the manual mode the keyer provides 8 to 50-wpm paddle or squeeze keying with dot memory, independent weight control and iambic character generation. A tap on the straight key of a Brown Brothers CTL combination key (or external push button) initiates the automatic program at the exact speed and weight used by the operator in the manual mode. At 20 wpm a full message sequence takes 15 seconds. The sequence terminates either automatically or by a tap on the dash paddle.

The *Mnemonic* keyer will operate both grid-block and cathode keyed transmitters. Power supply, cw monitor and speaker are built-in. All cables and connectors are provided.

The EK-39M Mnemonic keyer is available from dealers or directly from the manufacturer, and is priced at $179.95 less the individually tailored plug-in memory which is ordered direct from the factory per user message instructions. The memory is priced at $59.95. For more information, use *check-off* on page 88, or write to Curtis Electro Devices, Box 4090, Mountain View, California 94040.

## amateur radio techniques

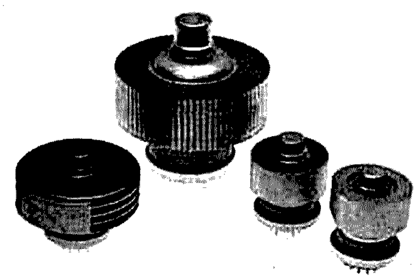
The new enlarged edition of this very popular book is now available, and the author, Pat Hawker, G3VA, has included many new topics. The emphasis is on solid state, with many new semiconductor circuits for amateur equipment. Chapters are included on semiconductors, components and construction, receivers, oscillators transmitters, audio and modulation, power supplies, antennas and troubleshooting and test equipment. This is an excellent book that presents many unique circuits and construction techniques, as well as solutions to old problems.

The chapter on receivers, for example, discusses modern communications receiver design, including front ends, gain distribution, mixers, oscillators, i-f stages, spurious responses, crystal filters, detectors, noise limiters, cross modulation, and direct-conversion systems. All the latest techniques are there, including many that were developed for military and commercial use, and have yet to filter down to amateur gear.

The other chapters are equally as diversified as the one on receivers, and provide a broad look at modern communications circuits and systems and how they may be used by the amateur. 208 pages. 470 diagrams. $3.50 from Comtec Book Division, Box 592, Amherst, New Hampshire 03031

## new zero bias triodes

Eimac has just announced a new family of ceramic-metal zero-bias triodes that are suitable for use up to 450 MHz



or so. Three versions are featured: the 8875 transverse air cooled, the 8873 conduction cooled and the 8874 axially air cooled. The 8875 is a high-mu power triode having 300 watts plate dissipation and capable of 1200 watts peak in Intermittent Voice Service. The 8874 features 400 watts plate dissipation. In the January, 1971 issue of *ham radio* we will present application articles on the 8875 and 8873.

The big tube in the center of the photograph is a new 1500-watt zero-bias triode, the 8877. In the February issue we'll have an article describing its application in a six-meter linear amplifier. Right now, your're not even supposed to be seeing it!

For more information on these new Eimac power tubes, use *check-off* on page 88, or write to Eimac Division of Varian, 301 Industrial Way, San Carlos, California 94070.

## high-power rf transistors

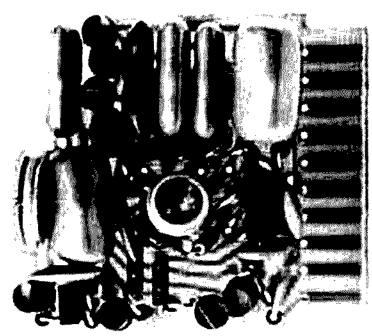
Two npn silicon rf power transistors for vhf power amplifier applications to 175 MHz are available from Motorola Semiconductor Products. The transistors, types MM1552 and MM1553, use balanced-emitter construction for extreme electrical ruggedness; load mismatch conditions of 10:1 at 75 watts output can be withstood by either device.

Intended for use as high-power class-C amplifiers in the 100-175 MHz frequency range, the MM1552 and MM1553 have the high peak power capabilities and breakdown voltages required for a-m applications. The MM1552, for example, is rated at 90 W peak power output for

18 W (maximum) input at 150 MHz (27 Vdc power supply), permitting close to 100 percent modulation with a 25 W carrier at $V_{CC}$ = 13.5 Vdc. For the MM1553, the same figures are obtained with a 13.5 W (maximum) input. In fm or cw service, both devices are capable of a continuous 75 watts output at 150 MHz.

Much of the fine broadband performance of the transistors comes from their low lead-inductance, strip-line package — Motorola Case 145C-01. For more information use *check off* on page 88, or contact the Technical Information Center, Box 20912, Phoenix, Arizona 85036.

## tone encoder/decoder



Because of the increased demand for a reliable all solid-state sub-audible continuous-tone encoder/decoder that is small enough to fit internally in small two-way radios, Alpha Electronic Services developed the SS-80H. Measuring only 1½ x 1¼ x ½ inch, it is probably the smallest unit of its kind. The SS-80H meets or exceeds all EIA specifications, and when coupled with the TN-91H frequency-determining module can be easily installed in equipment where space is a problem. The SS-80H is completely compatible with private line, channel guard or other standard frequency tone-quieting devices. It is also available with special tone frequencies, allowing greater use of congested channels. The SS-80H uses no mechanical reeds. For more information, use *check-off* on page 88, or write to Alpha Electronic Services, Inc., 8431 Monroe Avenue, Stanton, CA. 90680.

# short circuits

## ST-5 demodulator

Several component values were omitted from the discriminator section of the schematic on page 14 of the September issue. These values are determined by the shift, and are shown in the following table:

| shift | 2125 – 2975 | 1275 – 2125 |
|---|---|---|
| R1 | 4700 | 1500 |
| R2 | 33k | 8200 |
| R3 | 5600 | 2200 |
| R4 | 91k | 68k |
| C1 | .068 | .18 |
| C2 | .033 | .068 |

## frequency counter

In fig. 5, page 23 of the July issue, IC2 should be shown as an MC799P. In fig. 2, page 19, the 20-pF capacitor in the crystal-oscillator circuit should be an N750 type for temperature compensation. In fig. 7 the 170-volt power supply output is shorted – this should be shown as a shielded line to ground. Also in fig. 7 one of the switch contacts is labeled *control module pin 14;* this should be *clock module pin 14.* And finally, in fig. 5, an electrolytic capacitor on the bottom right side of the drawing is shown with the wrong polarity.

## wwv receiver

On page 68 of the July issue the schematic for the simple WWV receiver has no detector stage – a diode should be installed between the 1.2k and 20k resistors on the output of the SA21 integrated circuit, cathode end toward the 20k resistor.

## microwave hybrids

On page 61 of the July issue, under applications, the sentence that states, "A detector-indicator, such as a receiver with an S-meter connected at port 3 . . ." should say port 1. If imbalance is measured at port 3 the input signal must be applied to port 1, in which case the signals at ports 2 and 4 will be 180 degrees out of the phase, and not suitable for balancing antenna sections. The situation is correctly described in the previous paragraphs.

# ham radio

## index to volume III – 1970

This index covers all articles published in **ham radio** during 1970. The articles are listed alphabetically under each category along with the author, page number and month. Categories are: antennas and transmission lines; commercial equipment; construction techniques; fm and repeaters; integrated circuits; keying; measurements and test equipment; miscellaneous technical; power supplies; propagation; receivers and converters; RTTY; semiconductors; single sideband; transmitters and power amplifiers; and vhf and microwave. Articles followed by (HN) appeared in the **ham notebook.**

### antennas and transmission lines



### audio

## miscellaneous technical



## power supplies

## receivers and converters



## RTTY



## semiconductors

## single sideband



## transmitters and power amplifiers



## troubleshooting



## vhf and microwave